WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

JULY 1954

VOL. 31 No. 7 THREE SHILLINGS AND SIXPENCE

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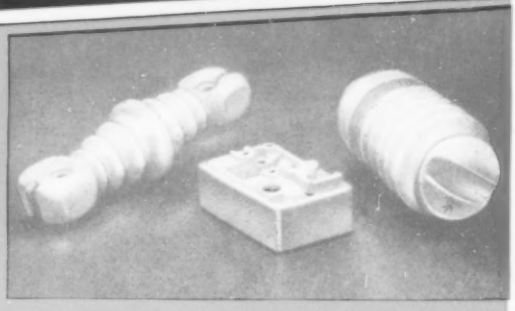

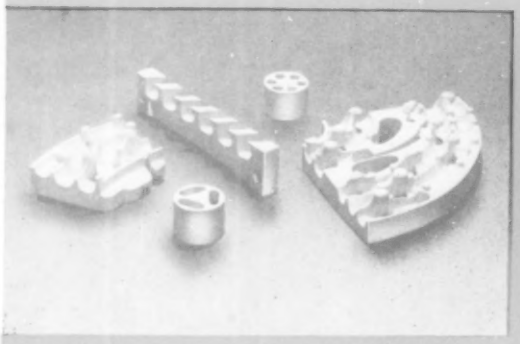

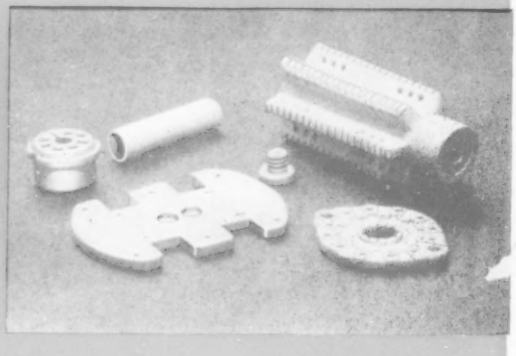

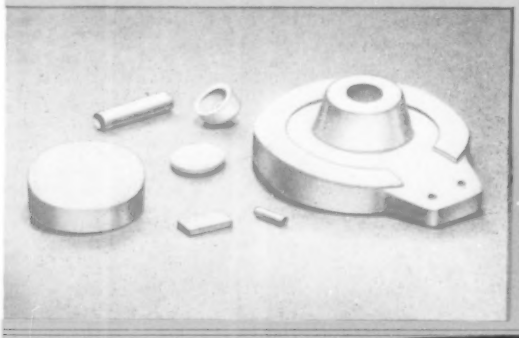

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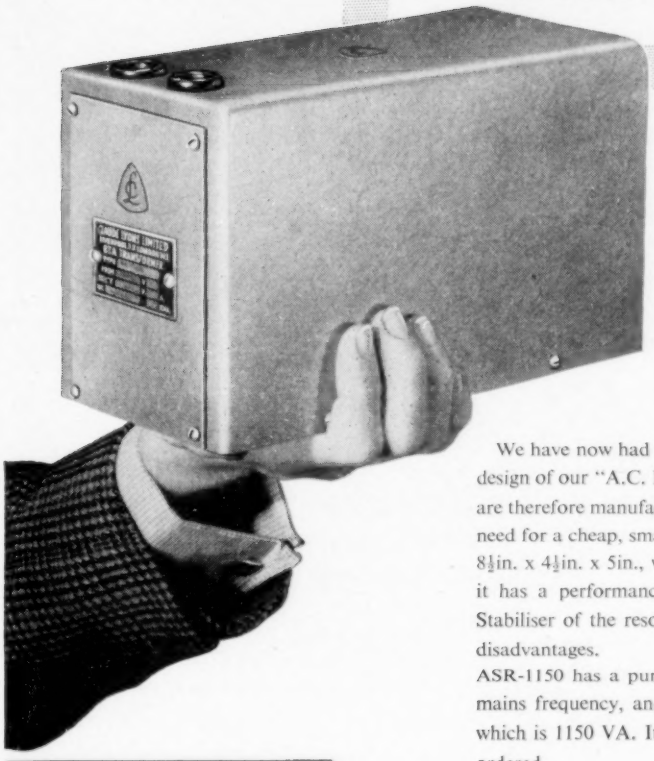
BL12D

A NEW-PRINCIPLE A-C AUTOMATIC VOLTAGE STABILISER



We have now had released to us by a Government Department the design of our "A.C. Mains Regulator, Automatic, Step, Mark II", and are therefore manufacturing this unit for general sale. It ideally fills the need for a cheap, small and light Stabiliser. Although it measures only 8½in. x 4½in. x 5in., weighs as little as 11 lbs., and costs only £24 net, it has a performance fully equal to any similarly rated Automatic Stabiliser of the resonated, saturated core type, without any of the disadvantages.

ASR-1150 has a pure output waveform, is unaffected by changes in mains frequency, and works equally well from no-load to full load, which is 1150 VA. It has a stabilised output at 230V unless otherwise ordered.

Many other Automatic Voltage Stabilisers are now manufactured by us, and all are available for immediate delivery. In some cases the constancy of output is as high as 0.15%. Models are available from 200 VA to 30 kVA, single phase. 3-Phase Stabilisers are also available. Prices are *extremely* competitive.

We can supply from stock all types of American tubes, condensers, valves, potentiometers, etc.

The *NEW "ASR-1150"* costs only *£24* net

Complete information is obtainable from:

CLAUDE LYONS LTD., STABILISER DIVISION,
QUEEN'S HOUSE, 180-182 TOTTENHAM COURT ROAD, LONDON, W.1
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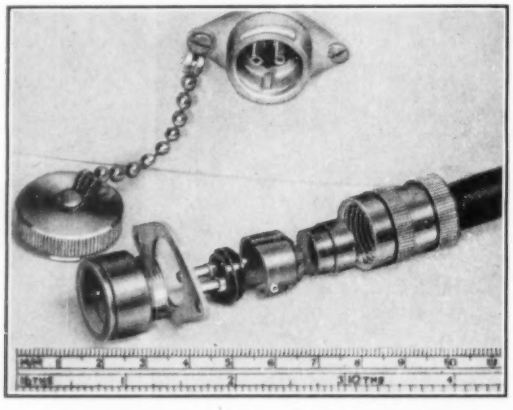
MEMO: If you are interested in infinitely-variable Transformers, do not forget the almost indispensable "VARIAC" (Reg'd. Trademark). Models are available from 170VA to 21kVA. Our Catalogue V-549 (3rd Edition) tells the whole story, and will gladly be mailed free and post free, on request.

A

"SCREENECTOR" REGD.

Lightweight plugs and sockets for coaxial, 2 or 3-pole assemblies

- **Versatile**
- **Perfect contact**
- **High insulation resistance**

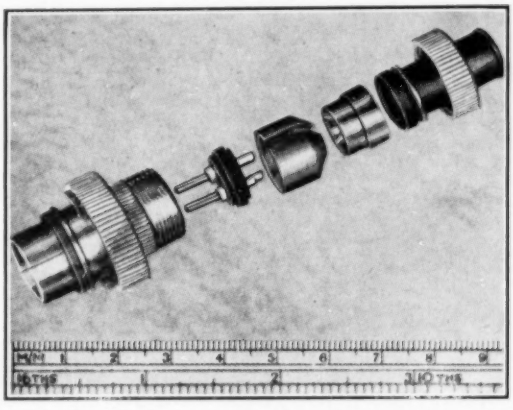


L 6 2 5 / C A dust cover, with the "quick" thread locking device, has been developed for use with these connectors and is supplied complete with chain attachment.

Assembly	Coaxial	Twin	3-pole
Free plug	L722/P	L625/P	L715/P
Fixed socket	L722/S	L625/S	L715/S
Adaptor bulkhead	L723	L689	L716
Free socket	L724	L690	L717

Type	Characteristic impedance, ohms*	Contact resistance at 1 amp.	*Capacitance P. & S. mated	
			Conductor/ conductor	Conductor/ screen
Coaxial	75	(less than	—	1.9 pF
2-pole	100	2 milli-	1.4 pF	2.3 pF
3-pole	—	ohms ea.)	2.0 pF	2.7 pF

Approx. at 1 Mc/s.



THIS range of light weight screened connectors is based upon the draft R.E.C.M.F. specification for a non-reversible screened twin plug and socket to load cables up to 0.24 in. diameter over the braid. At the same time, the design has been made versatile so that single and multiple contacts may be assembled into the common housing. This housing is machined and therefore possesses an appearance in keeping with modern instrument finish.

Although a "quick" thread locking ring forms part of the assembly, the screen skirt is resilient, thus maintaining excellent contact should the locking ring be inadvertently left loose.

The contact assembly insulator is produced from a moisture resistant nylon-filled phenolic moulding material, and the design includes lengthened leakage paths between contacts in order to maintain high insulation resistance which is 0.6×10^5 MΩ at 500 V. between pins or pins and frame.

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A New Portable Wave Analyser

THE WAYNE KERR MODEL A.321

An audio frequency analyser of the heterodyne type to measure the relative levels of the components of a complex waveform down to −75 db below fundamental.
It is light, compact and simple to operate and can be supplied either in a transportable case or for standard 19" rack mounting.

Specification

FREQUENCY COVERAGE:	*50—20,000 c/s in two ranges*
FREQUENCY STABILITY:	*Better than ±1%*
INPUT LEVEL:	*Between −20 and +20 db relative to 1mV into 600 ohms adjusted by a 5 step attenuator and fine control of 0—10 db*
FREQUENCY RESPONSE:	*Flat to within ±1 db over the whole working range.*
LEVEL STABILITY:	*±1 db if mains supply voltage remains within ±5%*
ACCURACY OF HARMONIC MEASUREMENTS:	*Harmonics below fundamental* *to −55 db ±5%* *to −75 db ±10%* *or ±1 db whichever is the greater.*
INPUT IMPEDANCE:	*100,000 ohm unbalanced (greater than 25,000 ohm, balanced, can be supplied to order)*
SELECTIVITY:	*Constant over the frequency range. 2 stage crystal filter, mid band frequency 100 Kc/s. Flat top (±1 db) pass band extending over 10 c/s ±1 c/s.*
HUM LEVEL:	*−55 db*
POWER SUPPLY:	*110/115 and 200/250 volts 40/60 c/s*
POWER CONSUMPTION:	*Approx. 60 watts*
DIMENSIONS:	*In case 20" × 9" × 8½"*
WEIGHT:	*31 lbs. approx.*

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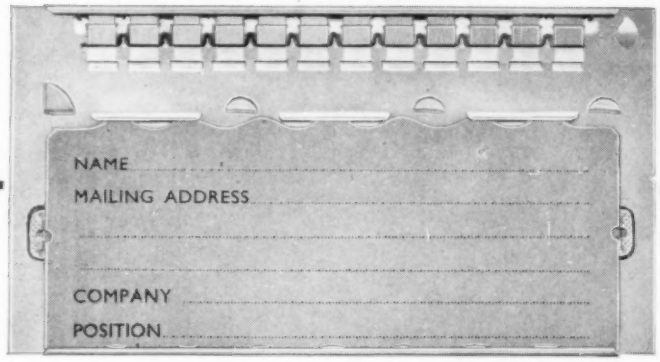

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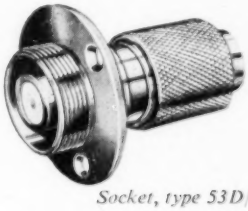
co-axial cable connectors



Socket, type 53D/20M

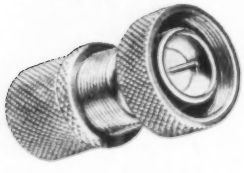
Illustrated is a range of Telconnectors which enable small flexible cables to be readily attached to larger feeder cables. In all cases the Socket Termination is designed to accept the standard 53P series Plug. A further step-down in size is possible by using a 53/58 Adaptor, the female section of which will accept the standard 58P series Plug.

Adaptor, type 53/58

All types are readily available in various sizes and combinations. Write for publication T/1.

TELCON RF cables

Plug, type 53P/29M



Protective Cap, type 53S

THE TELEGRAPH CONSTRUCTION & MAINTENANCE CO. LTD
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weight 0·42 grammes

size

Length 3·2 mm

Diameter 7·2 mm

SenTerCel Types M1 and M3 rectifiers are low in cost and offer many advantages. They replace equivalent thermionic valves and can be wired directly into circuit ; wiring is reduced and valve-holders are eliminated.

Both types operate at minimum input levels of 0.5 volts, type M1 at frequencies up to 5 Mc/s and type M3 up to 100 kc/s.

APPLICATIONS

AGC rectifiers : muting circuits : contrast expansion and compression : level indicators : modulation depth indicators : limiters : automatic frequency control.

Type M1

Average Characteristics

Self Capacitance	22 pF
Forward Resistance at 5 V D.C.	10 kΩ
Reverse resistance at 5 V D.C.	1,000 MΩ
Maximum Peak Inverse Voltage	68 V
Minimum A.C. Input	0.5 V
Maximum Frequency	5 Mc/s.

Type M3

Average Characteristics

Self Capacitance	65 pF
Forward Resistance at 5 V D.C.	1.2 kΩ
Reverse Resistance at 5 V D.C.	45 MΩ
Maximum Peak Inverse Voltage	68 V
Minimum A.C. Input	0.5 V
Maximum Frequency	100 kc/s

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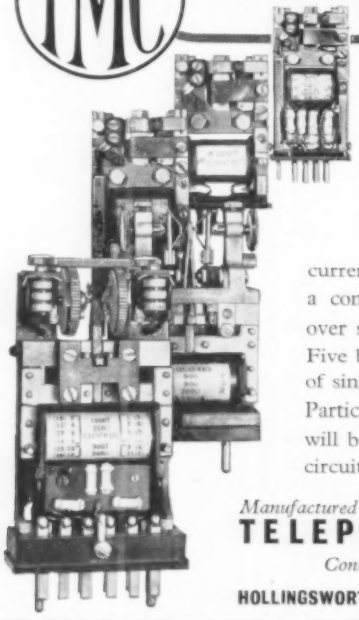
TMC

CARPENTER POLARIZED RELAYS

—have these outstanding features



The Carpenter Polarized Relay will respond to weak, ill-defined or short-duration impulses of differing polarity, or it will follow weak alternating current inputs of high frequencies and so provide a continuously operating symmetrical change-over switch between two different sources.
Five basic types are available with a wide range of single and multiple windings.
Particulars of the type best suited to your purpose will be gladly supplied if you will send us your circuit details.

- High operational speed
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Photo by courtesy of Telegraph Condenser Company Ltd.

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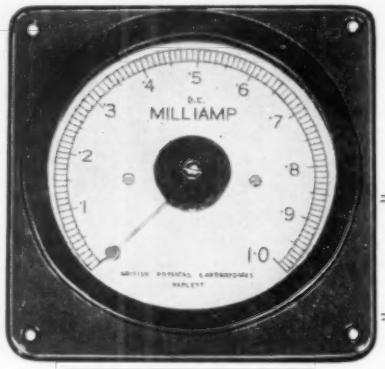

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SCALE LENGTH: 9″
RANGE FROM 0-500μA

The meter with the dead beat movement

The meter measures only 5″ across and the depth behind the panel is only 2″ yet it retains a scale length of 9″. It is available with spade or knife-edge pointer and special scales can be supplied to customer's specification if required. Send for prices and full details.

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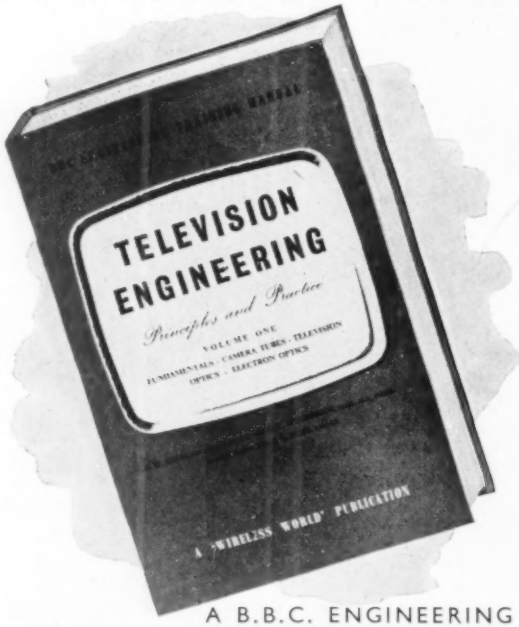

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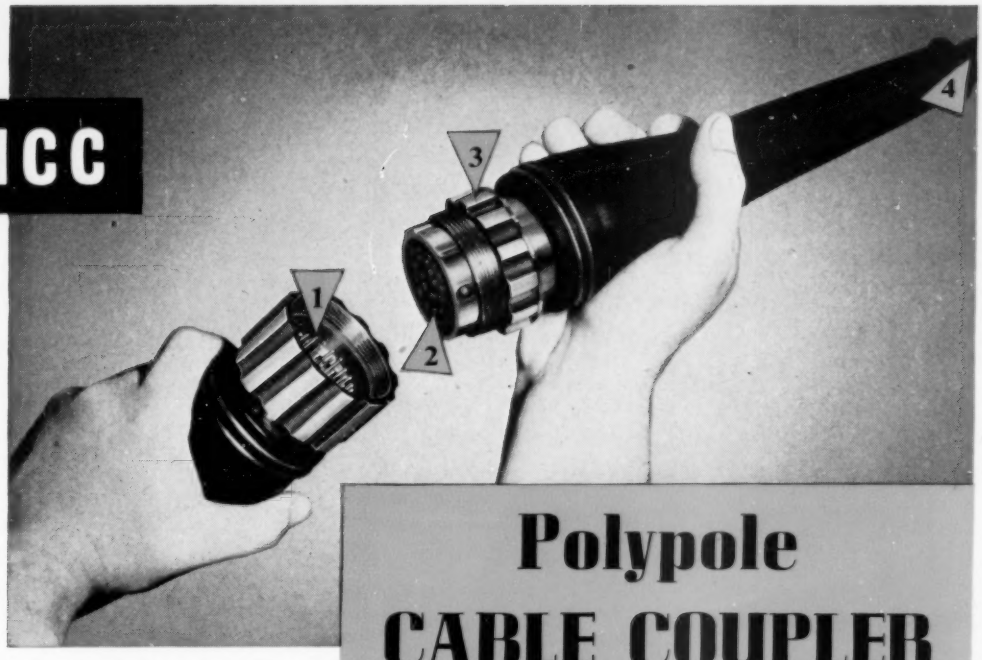

Polypole CABLE COUPLER

For- *specialised remote control—centimetre radio links—ground radar—outside broadcast television*

BICC Couplers and Cables are intended for the outdoor inter-connection of equipment, such as that mentioned above. Each application calls for composite trailing cables containing both R.F. units and other polythene insulated conductors.

BICC Polypole Mark III Couplers are available in two versions, designed for use with two standard types of BICC outdoor trailing cables. The Mark IIIA cable and coupler incorporates three coaxial circuits, and the Mark IIIB three screened twin circuits. In addition, both cables contain three triplets and 21 other conductors.

The couplers are permanently moulded to the ends of the cable in the factory. This technique provides a remarkably robust coupler which is virtually free from the hazards of conductor breakages near to, or within the coupler.

If you are interested in the uses of BICC Polypole Cable Couplers, we will be pleased to send you further information.

Note these important features

1. The couplers are assembled with the conductors in tension to ensure that they each contribute their share of the total strength.
2. Polythene injection moulding permits a watertight assembly.
3. Screwed lock rings provide forced engagement and withdrawal. The overall metal housing can also be easily replaced should it become damaged.
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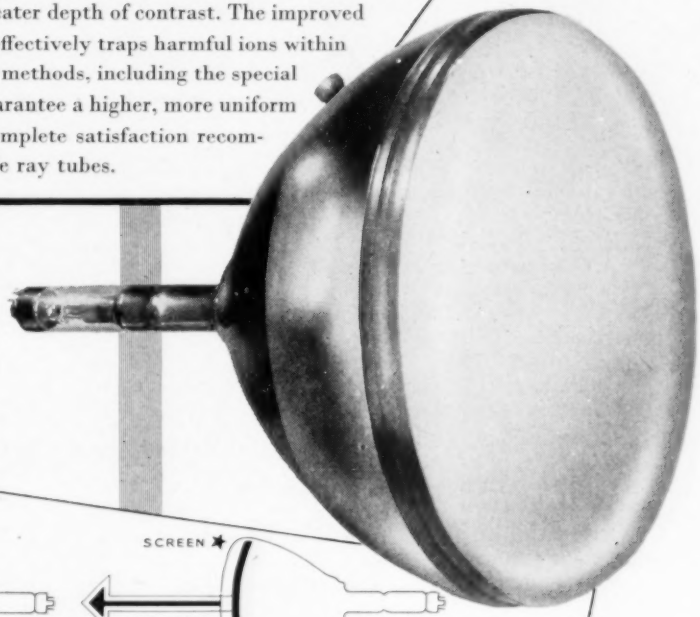
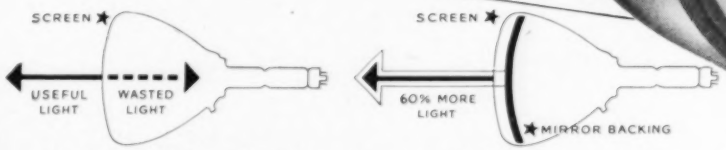
You cannot beat the new Ediswan Mazda Aluminized Cathode Ray Tubes for picture quality or length of life. Ediswan Aluminizing gives 60% brighter pictures with greater depth of contrast. The improved ion trap tetrode electron gun effectively traps harmful ions within the tube. Ediswan production methods, including the special in-line exhausting process, guarantee a higher, more uniform standard of efficiency. For complete satisfaction recommend Ediswan Mazda cathode ray tubes.





Without Aluminizing

Without Aluminizing, tubes waste half their light by backward reflection. To counteract this the brilliance must be increased which shortens tube life.

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A compact, sensitive, yet robustly suspended galvanometer has been developed in our laboratories and, having undergone extensive trials, has proved to be extremely reliable for all purposes. All galvanometers are fitted with shock-absorbing non-sticking stops, thus overloads of up to 100 times full-scale current may be applied without risk of damage to the movement.

It is available as a compact unit for bench use, mounted in a bakelite case measuring only 12 × 8 × 4 cms. and is eminently suitable for mounting in the client's own equipment—a simple bracket mounting is incorporated to facilitate this.

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Periodic Time: 2 seconds.
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WIRELESS ENGINEER

Vol. 31 JULY 1954 No. 7

Improved Morphology of Electromagnetics

THIS is the title of an article of 83 pages in the *Bulletin of the Technical University of Istanbul* (1952) by Paul de Belatini of the Telecommunication Department. The dictionary tells us that morphology is the science of organic form. There has been a tendency in recent years to merge the two systems, one based on the unit charge and the other on the unit pole, into a common system, but the author suggests eliminating the anomalies of the usual duality of electricity and magnetism by "replacing the duality by triplicity, where the third sector is that of 'dielectricity' in duality to magnetism, with electricity midway between them".

The meaning of this is made clear by the following:—

TABLE II

Triple Analogies

$\mu = B/H$	$\kappa = J/E_c$	$\epsilon = D/E_c$
$B = \Phi/F$	$J = i/F$	$D = \Psi/F$
$W_k = \frac{1}{2}M\Phi$	$P = ei$	$W = \frac{1}{2}eq$
$D_k = \frac{1}{2}HB$	$N = E_cJ$	$D_p = \frac{1}{2}E_cD$

Double Analogies

$H = M/l$	$E_c = e/l$
$M = \Phi R_M$	$e = iR_E$
$R_M = l/\mu F$	$R_E = l/\kappa F$

Without Analogies

$C = \epsilon F/l$, $L = n^2/R_M$, $M = ni$, $q = eC$, etc.

In these formulae F = area, E_c = electric field strength, J = current density, W = energy, D = energy density, also displacement density P = power, N = power density, e = e.m.f., M = m.m.f., R_e = resistance, R_m = reluctance, κ = conductivity. We have copied the author's symbols except where they were obviously wrong.

In the triple analogies the first column is magnetic, the third electrostatic or dielectric and the second column is what the author regards as their electric analogues. In the double analogies one of the three is missing from ordinary usage, and there are many formulae without any analogy. The author's object is to fill these gaps as far as possible by introducing new conceptions and reshaping equations. He claims that this results in a perfectly symmetrical classification of electromagnetic quantities and laws. A tabulation is proposed, which he calls the Periodic System, "in which all known quantities are included and the laws are shown by the geometry of the tabulation". This is applied to other branches of physics, and he develops three periodic systems, viz., electromagnetic, mechanical and thermal, which he then unifies in a single general periodic system, with three absolute basic dimensions, length, time, and energy (or one of its time derivatives) and one relative dimension, viz., "the impedance of the domain in question". The name 'periodic' is adopted from Mendeleev's chemical system.

An interesting example of the author's procedure is the introduction of the multi-plate-pair capacitor. The reciprocal of the capacitance is called the dielectric reluctance R_d and in the formula $R_d = l/(\epsilon F)$, l and F are regarded as the length and cross-section of the dielectric core, just as in the formula $L = 1/R_m$ for the inductance of a single turn inductor, R_m is the

reluctance of the core. If the inductor has n turns $L = n^2/R_m$, and to obtain the dielectric analogy, the author uses a number of capacitors in series but all using the same slab of dielectric as shown in Fig. 1. If the reluctance R_d of the whole slab is thus divided between p pairs of plates, the reluctance of each section will be pR_d and the capacitance of the p pairs in series will be

$$C = (1/p)^2/R_D$$

which is analogous to

$$L = n^2/R_M$$

The dual of n is $1/p$ because ωL and ωC are of an inverse character. This example shows how, by departing somewhat from ordinary usage, analogies can be devised and the gaps in the table filled up.

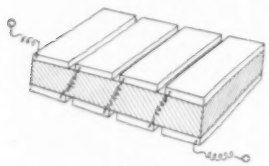

Fig. 1

His first example of the lack of analogy between magnetic and electric formulae is rather surprising; he says that although $\int_0 H dl = M$ where M is the m.m.f., the line integral $\int_0 F_c dl = 0$ where E_c is the electric field strength. This is surely wrong, since the line integral of E_c in a closed path is the e.m.f. which he designates e. He returns to the subject later on and it is seen that the anomaly is of his own invention. In a section entitled "Reshaping of the Maxwell vector equations" he uses three different symbols for electric field strength, viz., E, E_c and U; he says, "Up to now we have used a common space derivation, $E_c = de/dl$. Obviously we must split this into dielectric field strength $U = du/dl$ and the rate of voltage drop $E = de/dl$." The word 'obviously' is ominous. Four lines after stating that $\int_0 E_c dl = 0$, he has curl $E_c = -dB/dt$ which seems contradictory. He then says, "U is restricted to media characterized by ϵ or dielectrics; for conducting media characterized by κ it is the rate of voltage drop which is to be used." He does not say what is to be done when current flows in a conductor without any voltage drop. Consider two closed rings, one of copper and the other of a dielectric, on an iron core through which the magnetic flux is changing. The e.m.f. e induced will be the same in each ring, viz., $-d\Phi/dt$ and the electric field strength E in each case will be the e.m.f. divided by the length of the path around the ring. In the copper ring the current will be e/R and the current density $J = E/\rho$; in the dielectric ring the displacement density $D = E\epsilon$ where ϵ is the permittivity of the material. In neither case does potential enter into the case since all points of the ring are at the same potential; in both cases the effect is due to the induced electric field and not to fall of potential. In a recent programme for a meeting of the Education Discussion Circle of the Institution of Electrical Engineers we were surprised to see that one of the questions to be discussed was "Is there any electric field strength in a short-circuited ring?" One of the free electrons might be consulted as to why it was moving. Perhaps further light is thrown on his point of view by considering a dynamo and its external circuit. In the external circuit the electric field strength at any point is equal to the fall of potential per unit length, but in the armature winding there are three electric forces, viz., the induced electric field due to the movement in the magnetic field, the opposing electric field due to the fall of potential and the resultant electric field equal to the difference between them. The only one of these for which the line integral around the circuit is zero is that due to the fall of potential, and the author must have had this in mind when he wrote $\int_0 E_c dl = 0$, although this does not agree with curl $E_c = -dB/dt$ which he writes four lines further on. Since the line integral around the circuit of the forces due to the potential difference must be zero, the integral of the actual resultant force must be the same as that of the induced electric force; i.e., $-n\,d\Phi/dt$.

On page 42 he states that E = electric rate of voltage-drop and U = dielectric field-strength, and on page 43 we are told that $U = -E$ and curl $E = dB/dt$ and then on page 44 that curl $E = -dB/dt$. Apart from this final contradiction, surely the field-strength or electric force in a dielectric is in the same direction as the voltage drop.

We have only touched on a fraction of this lengthy paper, but we must confess that, after struggling with these apparent contradictions for many hours, we gave it up. In view of the great amount of work that has evidently been put into the paper, it is to be hoped that the author will make an effort to set out more clearly the initial basic sections.

G. W. O. H.

A Heaviside-Lorentz Network Theorem

IN both the April and May Editorials we referred to a theorem of Heaviside which was developed by Lorentz and recently discussed in a paper by van der Pol and also in a paper by Tellegen. In both cases the reference given was to Heaviside's Electrical Papers, Vol. II, p. 412. Our attention has been drawn to the fact that Heaviside is here merely quoting a special application of a general theorem which he had derived in Vol. I, p. 462, where a section is entitled "Work done by impressed forces during transient states".

G. W. O. H.

SELF-HEATING TRIODE FOR VOLTAGE STABILIZATION

By E. G. Hopkins, B.E., A.M.I.E.(Aust.).

(*School of Electrical Engineering, N.S.W. University of Technology, Sydney.*)

SUMMARY.—A self-heating triode for use in voltage-stabilizing circuits is described. It contains two planar oxide cathodes separated by a planar grid and acts as a variable resistance directly in series with the a.c. supply to small fixed loads. The valve has two auxiliary anodes which connect the grid circuit to the appropriate cathode during each half-cycle. A large cathode area of 8·4 cm² relative to the total power dissipation of 16 W renders the triode more suitable than conventional valves for operation in low-voltage circuits.

1. Introduction

THE alternating voltage supplied to a piece of small electrical equipment is sometimes stabilized or controlled by a pair of high-vacuum triodes or pentodes acting as a variable resistance.[1] The valves are usually connected across the centre-tapped secondary of a transformer whose primary is placed in series with the supply line. The transformer is necessary to increase the low equivalent conductance and the low rated current of standard receiving valves to more appropriate values.

The present paper describes a self-heating triode[2, 3] designed especially to replace the two valves and the transformer by acting as a variable a.c. resistance directly in the supply line to a constant or nearly-constant load. The triode has two parallel planar cathodes equally spaced about a planar grid. Each cathode acts alternately as anode and cathode. Heat generated by electron bombardment maintains the cathode temperature within its working range. The cathodes are provided with heaters which are used only for starting. Two auxiliary anodes clamp the grid circuit to the appropriate cathode during each half-cycle of operation.

The basic principles of self-heating valves are discussed in other papers.[2, 4]

2. Description of the Valve

The working parts of the triode are illustrated in Fig. 1. The cathodes are 42 mm in length and have a hollow rectangular cross-section measuring 10 mm × 1·5 mm outside, with a wall thickness of 0·25 mm. They are made of polished Grade A nickel and are supported at each end in the mica spacers by two nickel rods 1 mm in diameter. The cathodes are induction-heated during exhaust. One face of each cathode is completely coated with barium-strontium oxide. A length of 1 cm at the centre of each reverse side is also coated to form the emitting surface of a clamping diode.

Each cathode is provided with a well-insulated 6·3-V, 0·6-ampere heater which is retained within the cathode between 0·85-mm diameter nickel rods. It is proposed to replace the present heaters in future valves by heaters rated at 0·2 ampere and 33 V which will be connected in place of the electron stream during the starting period, as is shown in the circuit of Fig. 2. Any value of heater power between the limits of about 6 W and 15 W is satisfactory for starting the triode.

The grid consists of molybdenum wires 0·175 mm in diameter welded to 1-mm nickel side rods. Each side rod carries four carbonized nickel cooling fins 10 mm long and 5 mm wide. The plain nickel clamping anodes measuring 1 cm × 1 cm are attached to two supporting members which are eyeletted to the mica spacers.

The spacing between cathodes is 0·8 mm. This relatively large spacing is used only to avoid the necessity for close mechanical tolerances.

The envelope is 29 mm in diameter and 85 mm in length. The envelope, stem, base and heaters are standard "GT" receiving-valve parts.

3. Performance of the Valve

Fig. 2 shows a circuit in which the self-heating triode is connected, together with the shunt

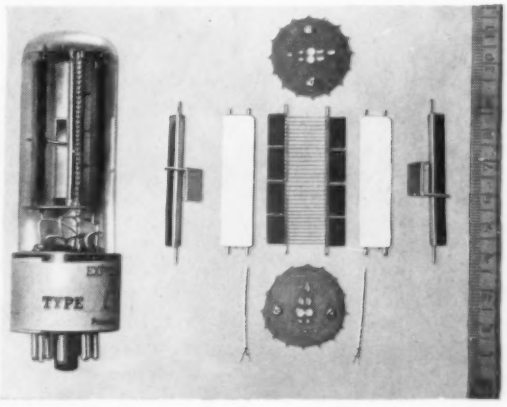

Fig. 1. *The parts of the self-heating triode.*

MS accepted by the Editor, September 1953

resistor R_s, between a constant load and an a.c. supply which varies about a nominal voltage of 240. Because the resistances R_g are large compared with those of the clamping diodes, each cathode is effectively connected to one end of the grid transformer winding during the half-cycle in which it is negative with respect to the other cathode. A control signal in phase with the anode voltage V_a is fed into the transformer so that the grid is driven negative with respect to the appropriate cathode by V_g, a voltage of sinusoidal form. The control signal, which is derived from a measuring circuit, maintains the load current, I_L constant as the supply voltage changes.

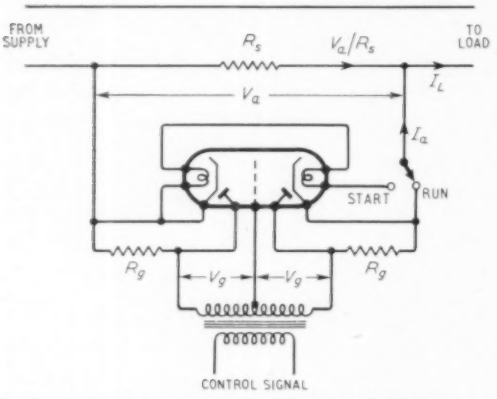

Fig. 2. *A circuit in which the self-heating triode acts as a variable resistance to maintain a constant load current, I_L when controlled by a mains-frequency signal derived from a measuring circuit.*

The characteristic curves of r.m.s. anode current, I_a, versus r.m.s. anode voltage, V_a, measured with sinusoidal anode and grid voltages are shown in Fig. 3. Since $I_L = I_a + V_a/R_s$, the operating point is defined by the intersection of the straight line $I_a = I_L - V_a/R_s$ and the V_a/I_a characteristic for the particular grid voltage being considered. Harmonic currents may be ignored.[2] The region between the curves of maximum and minimum dissipation (16 W and 10 W) defines the safe working range of the valve. The power delivered to a load of unity power factor is I_L (240 — V_a'), where V_a' is the centre value of anode voltage. R_s is chosen to give the best centre value and range of anode voltage with the specified load current. For example, when R_s is 400 ohms a control of ± 22 V is possible for a load current of 0·4 ampere (56 W load power), and when R_s is 600 ohms ± 40 V variation is possible for a load current of 0·32 ampere (40 W load power). The voltage-gain of these circuits is about unity and the total harmonic distortion introduced into the load current is about 2%.

4. Comments

The emitting areas of the cathodes are much greater than those of two conventional valves of similar total bulb dissipation. For example, if a double triode were designed with cathodes of similar area to those of the present self-heating triode it would require a filament power alone of at least 16 W, which equals the total bulb dissipation of the self-heating triode. This power would have to be dissipated by the anodes and bulb of such a valve in addition to the power lost at the anodes themselves.

The present triode will operate stably between the limits of 8-W and 25-W dissipation. The range 10 W to 16 W quoted above was chosen arbitrarily pending an investigation of life performance with relatively large numbers of valves.

With negative grid voltage, the convergence of the electron streams between grid wires causes a large increase in current density on the bombarded regions of the cathode surfaces. In the present experimental valves, which use sprayed coating and highly polished base metal, considerable localized erosion of cathode coating takes place. Experiments are continuing to ascertain whether the base metal surfaces[5, 6] and coatings, which have been developed to improve thermal and electrical conduction between cathode surface

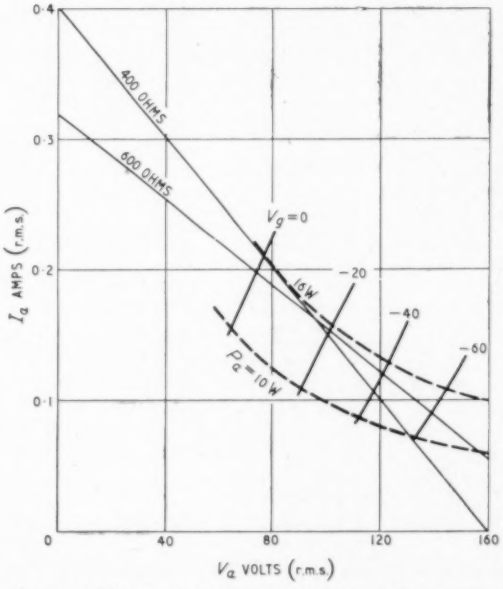

Fig. 3. *Characteristic curves of r.m.s. anode current versus r.m.s. anode voltage with sinusoidal anode and grid voltages. The load lines intersect the current axis at the load current I_L and have a slope equal to R_s, the shunt resistance of Fig. 2.*

and base metal, will reduce this effect to negligible proportions.

It is thought that the performance of the present valve may be improved by a small increase in cathode area and a decrease in inter-electrode spacing to accommodate a variation in V_a of ± 40 V about a centre value of 80 V for a load current of 0·5 ampere. This performance would be adequate for stabilizing the supply voltage to pieces of electronic equipment with power consumptions of up to 80 W or the heater voltage of a group of about 40 receiving valves.

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SPECTRUM OR WAVEFORM EQUALIZATION?

By D. A. Bell, M.A., Ph.D., M.I.E.E.

(*University of Birmingham*)

SUMMARY.—The relationship between the wave-filter approach to equalization and the approach via differentiating and integrating circuits is discussed in the light of their historical development and of present-day needs in both communication and servo systems.

1. Introduction

THE supremacy of the electric wave filter in circuit engineering is a phenomenon of the past quarter-century; and this, together with the general use of steady-state methods of circuit calculation, has tended to blind us to the possibility of carrying out engineering design in the time–amplitude plane instead of in the frequency plane. It was, therefore, very refreshing to read G. G. Gouriet's paper[1] (strangely entitled "*Spectrum* Equalization" when it approaches the problem from the angle of time functions rather than frequency functions) on the use of differentiating and integrating networks to restore the waveform of a signal which has been distorted at some point in the communication channel.* It is of course agreed that no system of correction can restore a component of the signal which has been *entirely* eliminated in transmission, or has been reduced to an intensity well below noise level: any process of equalization must, in principle, operate by levelling down to the weakest of the surviving components, accompanied usually by a uniform scalar amplification of the whole signal either before or after equalization. Moreover, if the specification for equalization requires that it should introduce no time delay, one can only work on the components as they exist at the first instant of the received signal.

An example may clarify this last point. If delay is permitted, the response of a typical physically-occurring low-pass filter can be phase-equalized by means of all-pass filter sections;[2] this will improve the symmetry and steepness of rise of the output response to a step-function input without introducing additional attenuation, but it will introduce delay. Alternatively, Bode has stated[3] that the frequency-discriminating effect of any physically-realizable minimum-phase network can be cancelled by adding in series with it an inverse network, the pair together then having a transfer characteristic of zero phase and constant magnitude. The apparent time delay (Gouriet's 'virtual delay') of the first circuit must then have been cancelled by the second circuit, but only at the cost of an overall attenuation not less than that at the critical point on the characteristic of the first network. (The 'critical point' may be defined either in terms of spectral response as the maximum frequency which is required to be transmitted, or in terms of waveform response as the earliest time at which full response to a step function is required.)

2. History

It is interesting now to note that an early proposal to use differentiating circuits for equalization related to the telegraph signals received through a submarine cable; i.e., it was concerned with counteracting the frequency-dispersion of a distributed circuit. This was described in U.S. Patent No. 1,315,539, issued to J. R. Carson on 9th September, 1919, the first claim being as follows:

"In a wave form correcting device, means

* The author has also had the pleasure and benefit of personal discussion of this topic with Mr. Gouriet.

MS accepted by the Editor, November 1953

for producing from the arrival current separate currents having wave forms corresponding substantially to the arrival current itself and one or more of its derivatives respectively, and means for superimposing the separate currents in a single output circuit."

Carson proposed in the first place to use inductive differentiating circuits (the low-inductance primary of a transformer was to be fed from a comparatively high-impedance source so as to carry a current proportional to the signal amplitude, and the secondary e.m.f. proportional to the derivative of the current was to be applied to a valve grid or other high-impedance circuit) with valves as buffer amplifiers between successive differentiations. Partial outputs could be tapped off along a chain of such differentiators to give a sum of the form

$$V_o = (1 + ap + bp^2 + cp^3 + \ldots.)V_i \quad .. \quad (1)$$

where V_o is the output of the equalizing system, V_i its input, and $p = d/dt$. So long as the differentiating circuits are nearly perfect and the various derivatives are combined by simple addition, such a system has the advantage that the several coefficients $a, b, c \ldots.$ in equation (1) can be adjusted separately. But Carson also suggested taking the output simply from the end of a cascade of imperfect differentiators. The output would then be of the form

$$V_o = (1 + a_1 p)(1 + a_2 p)(1 + a_3 p)\ldots. (1 + a_n p) V_i \quad .. \quad .. \quad .. \quad .. \quad (2)$$

and the coefficients of the powers of p in the polynomial form of the right-hand side of (2) would not be mutually independent.

Formula (2) is in fact a typical circuit *frequency response*, in which p is interchangeable with $j\omega$. The transition from the 'derivative' interpretation to the 'filter' interpretation is carried a stage further by considering American patent No. 1,586,821 issued to R. C. Mathes on 1st June, 1926.* Mathes placed far less stress on the use of buffer amplifiers and the desirability of mutual independence of the several differentiating circuits. He also proposed quite complex circuits, ranging from Fig. 1(a) to (b), as the 'differentiating' units, though these clearly would not produce pure single derivatives. When one has circuits as complex as this with, in some cases, mutual interaction between successive stages, one would nowadays design them in terms of frequency response as filters; the mathematical theory of electric wave filters barely existed in 1919, but has since become highly developed.

* Carson's patent, dated 9th September, 1919, was based on his application filed on 3rd July, 1918. Mathes's application was originally filed on 19th June, 1918 (i.e., a few days *before* Carson's), but was left in abeyance until 27th March, 1925, so that Mathes's patent was not granted until 1926, seven years after Carson's.

3. Application to Communication Systems

Now Gouriet proposes[1] to abandon the idea of equalization of the spectrum by wave filters, and to design instead in terms of equalization of the system response by means of differentiating circuits. Two questions are inevitable: "Is this really different?" and "Does it give any substantial advantage?" So far as the ultimate theory is concerned, there is no difference. The loose statement above that p and $j\omega$ are interchangeable can for this purpose be made sufficiently rigorous by pointing out that the various forms of operational calculus allow the time-function response to be completely derived from a frequency (p or $j\omega$) specification of the circuit. Hence correction of the $j\omega$ response necessarily corrects the waveform response and vice versa.*

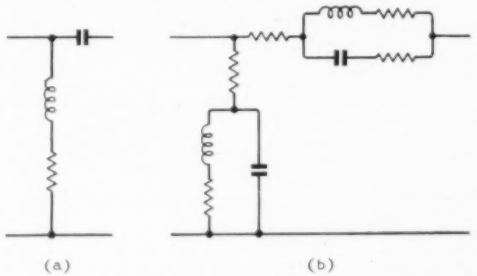

Fig. 1. *Complex differentiating circuits proposed by Mathes.*

The practical advantage of derivative equalization, however, is that it greatly simplifies the construction of an approximate correcting circuit for the overall distortion which may have been contributed to by many separate circuits of unknown form. The theoretical method of equalizing by networks might be to factorize the total transfer function so that its inverse could be constructed from a cascade of networks, using in principle one section of correcting network per factor in the transmission function; but this is cumbersome, and one would prefer to construct empirically a correcting network of polynomial form having the least number of terms consistent with a fair fit to the required characteristic. The great advantage of using pure derivatives is that each term in the polynomial can be adjusted *independently*, whereas a cascaded-network response of the type of formula (2) is such that change of any one parameter will affect the values of most of the terms in the polynomial and vice versa.

On these grounds it is difficult to see why the use of derivative equalization has been neglected from the time of Carson (1919) to Gouriet (1953).

* Only if the frequency response is expressed as a function of ω (instead of $j\omega$) will there be a possibility of failure through lack of phase agreement.

A possible answer is that whereas the wave filter is a passive network, and therefore has stable characteristics, the derivative equalizer commonly has a number of separate amplifiers, the relative gains of which affect the proportion of the different derivatives in the output; i.e., the equalization characteristic. This objection can now be largely overcome by the use of negative feedback to stabilize gains; e.g., by using cathode-followers as buffer stages between successive differentiating networks.

4. Application to Feedback and Servo Systems

For practical reasons, however, it is desirable to distinguish between communication systems and servo systems. In the former it can usually be assumed that although the communication channel is of limited bandwidth, the terminal equipment can be made with virtually infinite bandwidth; but in the servo system there is no such distinction between 'channel' and 'terminal equipment'. It is true that there may be a distinction between valve amplifying stages, of virtually infinite bandwidth, and rotary-machine amplifying stages of very limited bandwidth; but in this case the bandwidth limitation comes second in the chain, and one cannot put much pre-correction in the valve stages without reaching the power limitation which is the sole reason for the existence of the machine amplifying stages.

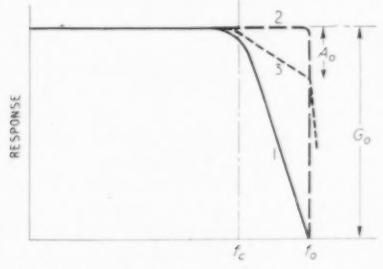

Fig. 2. *Gouriet's method of stabilization. Curve* 1 *is for the original characteristic and curve* 2 *for the equalized system. Curve* 3 *shows the final characteristic including a quadratic filter.*

In connection with feedback systems, Gouriet proposes an apparently new principle of stabilization: make the response perfect over a frequency range extending far beyond the required working band, then insert a quadratic low-pass filter to provide the cut-off. The quadratic filter cannot produce a phase-shift in excess of 180°, so the complete system is necessarily stable. In principle, however, this in no way differs from the Bode technique of adjusting the rate of cut-off of the system. According to Gouriet, stabilization is to be achieved as shown in Fig. 2: the original characteristic with nominal cut-off at f_c (curve 1) is equalized out to some higher frequency f_o (curve 2) and then a new and safe cut-off characteristic (curve 3) is introduced by means of a quadratic filter; i.e., the slope of curve 3 is less than 12 db per octave for a sufficient range.

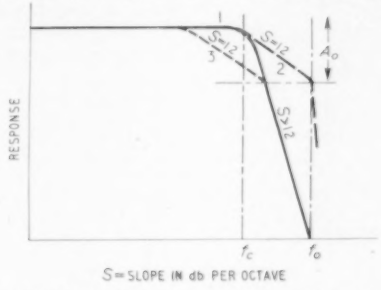

Fig. 3. *Bode's method of stabilization. S is the slope of the curve in db/octave.*

This clearly requires additional gain at high frequencies, the amount of which is arrived at as follows: (i) The additional gain G_o at f_o must equal the loss of the original circuit at f_o. (ii) The amount of attenuation A_o introduced by the quadratic filter at f_o must be at least equal to the loop gain, (to ensure stability when the overall characteristic leaves the quadratic characteristic 3 and begins to take on the large phase-shift which must be associated with the sharp cut-off at f_o) and this controls the distance of f_o from f_c. Now stabilization by the Bode type of technique is illustrated in Fig. 3. By increasing the bandwidth of a sufficient number of the stages, the open-loop cut-off is eased to 12 db/octave until the attenuation A_o exceeds the loop gain, after which the cut-off characteristic is allowed to take its own course (curve 2). Alternatively, if greater overall bandwidth is not obtainable one must ease the slope back as shown in curve 3, at the cost of curtailment of the working frequency band. The final result is then the same in Fig. 2 and Fig. 3—a cut-off slope reduced to a value not greater than 12 db/octave down to the point of zero loop gain. (Both methods ignore the question of shaping the characteristic within the working frequency band, so the final result may not after all be ideal for a servo system, but that is another matter.) However, the claim which may well be made for the system of Fig. 2 is that the best approximation to the equalization of curve 2 by means of derivatives can easily be established and verified empirically, and the quadratic filter is then simple to design and certain in operation. The design of the modifying filter for Fig. 3 is far more tricky, especially as calculated values of system characteristics can be unreliable in the presence of non-linearities.

Any equalization technique requires additional

amplification, whether this be included in series with the main signal chain and a network type of equalizer, or in parallel where the derivatives of the signal are amplified to the required extent. If there is no practical limitation and little economic limitation on increasing the amplification of the system at a fixed bandwidth, this presents no difficulty; but in a servo system real difficulties can arise.

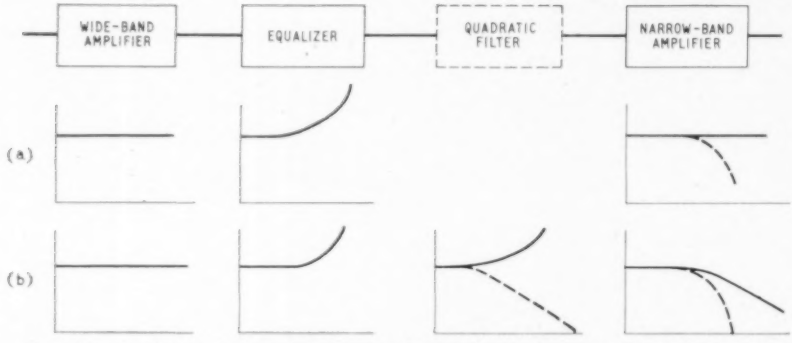

Fig. 4. *Transfer characteristic (a) for equalized system; (b) for equalized system plus quadratic filter. Cumulative characteristic shown broken line, unit characteristic shown solid line.*

In theory the gain-bandwidth product of a given type* of amplifier is independent of power rating, and in practice the gain-bandwidth may decrease only slowly as the power rating is increased. But as already remarked, there is for various reasons a change of type of amplifier from thermionic valve to rotary machine at a certain power level in most servo systems; and the bandwidth of the whole is effectively controlled by the machine stages. The achievement of curve 2 in Fig. 2 therefore implies a very large pre-distortion of the signal in the thermionic stages before it is applied to the machine stages, as shown diagrammatically at (a) in Fig. 4 for the exactly equalized system and at (b) with the addition of a quadratic cut-off. Even case (b) indicates a considerable rise in the peak power input to the narrow-band amplifier; and if the output power of the wide-band amplifier is fixed, the mean input to the narrow-band amplifier must necessarily be reduced, and its gain correspondingly increased. This will require more stages, and the question is whether one might not just as well have used the extra stages to shape the cut-off characteristic of the narrow-band amplifier, thus saving the equipment involved in the equalizer. As a first step one would in any case surely use part of the narrow-band amplifier as the quadratic filter, and this complicates the experimental setting-up of the equalizer.

There is certainly a case for adjusting the characteristics of a rotary amplifier by means other than a cascaded network, (e.g., by some form of feedback) because of the physical difficulties of constructing a network for extremely low frequencies; but this is not comparable with *equalization* by the addition of derivatives, since the latter requires additional gain with negligible time-constant (or in other words, additional gain at high frequencies) for the derivative components, and this additional gain is not available in a narrow-band amplifier. It is sometimes thought to be available when the amplifier output circuit has a much shorter time-constant than its input circuit.* But if the design of the amplifier or the load-matching conditions have been so adjusted that the input and output time-constants are approximately equal, there is in fact no such reserve.

* E.g., thermionic valve, rotary machine, magnetic amplifier.

5. Conclusions

The addition to the signal of derivatives amplified in parallel paths, instead of modification of the signal by frequency-selective networks connected in series with the main amplifying chain, greatly facilitates the empirical design of the correction for a given system; but the results which can, in principle, be obtained from the two systems (disregarding economic or constructional limitations) are identical, and the use of derivative circuits does not avoid the need for additional amplification at high frequencies (or with short time-constant, if one prefers to express it in that way). When using the derivative method, it must be remembered that the result depends on the constancy of the relative gains of any parallel paths through which different components of the output signal may be amplified.

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[2] A. E. Brain, "The Compensation for Phase Errors in Wide-band Video Amplifiers", *Proc. Instn elect. Engrs*, 1950, Vol. 97, Pt. III, p. 243.

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* Bode (ref. 3, p. 284) gives an example of the converse case, a cathode-feedback circuit in which the cathode-earth capacitance is assumed to provide the only time-constant.

L.F. COMPENSATION FOR VIDEO AMPLIFIERS

Part 1. Amplifiers without Feedback

By J. E. Flood, Ph.D., A.M.I.E.E.

SUMMARY.—The effect of distortion at low frequencies on the indicial response (i.e., the response to a unit-step input voltage) of amplifier circuits is considered. Low-frequency distortion causes the output waveform to depart from the ideal flat top of the unit step, but it can be made as flat as possible by making the maximum number of its differential coefficients to be zero at $t = 0$. If r derivatives are zero, the amplifier is said to have rth-order compensation. A general method for doing this is described and is applied to some typical resistance-coupled amplifier circuits. A single resistance-capacitance coupled stage can have second-order compensation and multi-stage amplifiers can have still higher orders of compensation. Experimental results confirm the analysis.

1. Introduction

VIDEO-FREQUENCY and pulse amplifiers must introduce as little change as possible in the waveforms of the signals they amplify. A useful criterion for judging the fidelity of an amplifier is its indicial response; that is, its response to a unit-step input voltage (i.e., a voltage which is zero when $t < 0$ and is unity when $t > 0$). The bandwidth of the amplifier is large, so the effects of gain and phase distortion at low frequencies and at high frequencies can be considered independently. Analyses of high-frequency distortion can be made from equivalent circuits which omit the coupling and decoupling networks. Analyses of low-frequency distortion can be made from equivalent circuits which omit stray capacitances and components included to compensate for their effect.

Distortion at high frequencies affects the initial build-up period of the indicial response and may cause overshoot. Distortion at low frequencies affects the subsequent waveform, causing it to depart from the ideal flat top of the unit step. The output waveform can be perfectly flat only if the frequency response is uniform down to zero frequency. If the amplifier cannot pass d.c. the output voltage ultimately must tend to zero with increasing time, but it is possible to design an amplifier so that the response may approximate to the ideal within given limits for a specified time.

The effects of high-frequency distortion have been fully dealt with elsewhere. In this paper the effect of distortion at low frequencies is analysed for some typical amplifier circuits. The indicial response is made as flat as possible by causing as many as possible of its differential coefficients to be zero at $t = 0$, as suggested by W. E. Thomson.[1] Some well-known results are included as particular cases.

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2. General Theory

The indicial response of the low-frequency equivalent circuit is given by its Heaviside operational expression $h(p)\mathbf{1}$ where

$$h(p) = \frac{p^n + a_1 p^{n-1} + a_2 p^{n-2} + \ldots + a_r p^{n-r} + \ldots + a_n}{p^n + b_1 p^{n-1} + b_2 p^{n-2} + \ldots + b_r p^{n-r} + \ldots + b_n} \quad \ldots \quad (1)$$

Expanding this in inverse powers of p gives

$$h(p) = 1 + \frac{c_1}{p} + \frac{c_2}{p^2} + \ldots + \frac{c_r}{p^r} + \ldots \quad \ldots \quad (2)$$

The values of the coefficients in this expansion are derived in Appendix 1 and are as follows:

$$c_1 = a_1 - b_1$$

$$c_2 = a_2 - b_2 - a_1 b_1 + b_1^2$$

$$c_3 = a_3 - b_3 - a_2 b_1 + 2 b_1 b_2 - a_1 b_2 + a_1 b_1^2 - b_1^3$$

. . . . . .

$$c_r = \sum_{s=0}^{r-1} d_s (a_{r-s} - b_{r-s}) \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

where $d_s = \sum \frac{(-1)^m m!}{\alpha!\,\beta!\,\ldots\,\epsilon!} b^{\alpha}{}_{\lambda} \cdot b^{\beta}{}_{\mu} \ldots b^{\epsilon}{}_{\rho} \quad (4)$

where the summation extends over all possible values of $\alpha, \beta \ldots \epsilon, \lambda, \mu, \ldots \rho$
and where

$$m = \alpha + \beta + \ldots + \rho; \quad \alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s;$$
$$1 \leqslant \lambda < \mu \ldots < \rho \leqslant s; \quad a_0 = b_0 = d_0 = 1$$

Now $$\frac{1}{p^r}\,1 = \frac{t^r}{r!}\,1$$

Therefore from equation (2):

$$h(t) = 1 + c_1 t + c_2 \frac{t^2}{2!} + c_3 \frac{t^3}{3!} + \ldots + c_r \frac{t^r}{r!} + \ldots \; (t > 0)$$

But by Maclaurin's theorem

$$h(t) = h(0) + h_1(0)t + h_2(0)\frac{t^2}{2!} + \ldots + h_r(0)\frac{t^r}{r!} + \ldots$$

where $h_r(0)$ is the rth differential coefficient of $h(t)$ at $t = 0$.

$$\therefore \quad h(0) = 1$$
$$h_1(0) = c_1\ ,\ h_2(0) = c_2$$
$$\ldots\ldots\ldots\ldots\ldots\ldots$$
$$h_r(0) = c_r\ ,\ \text{etc.}$$

The indicial response may be made to approach more and more nearly to the ideal [$h(t) = 1$ for all $t > 0$] by making more and more of the differential coefficients vanish at $t = 0$. Equation (3) shows that if $a_1 = b_1$, $a_2 = b_2$, . . . $a_r = b_r$, then $c_1 = c_2 = c_3 = \ldots = c_r = 0$ and the first r differential coefficients of $h(t)$ vanish at $t = 0$. The amplifier will then be said to have rth-order compensation of its indicial response. If the indicial response has $(n - 1)$th-order compensation, it will be called maximally flat. If $a_1 = b_1$, $a_2 = b_2$, . . . $a_n = b_n$, then equation (1) clearly gives a completely distortionless indicial response. It is shown in Section 5 that, except when $n = 2$, a maximally-flat indicial response corresponds to neither a maximally-flat gain–frequency response nor a maximally-flat phase–frequency response. This confirms that the gain–frequency response and phase–frequency response are not entirely satisfactory for judging the waveform distortion caused by gain and phase distortion at low frequencies.

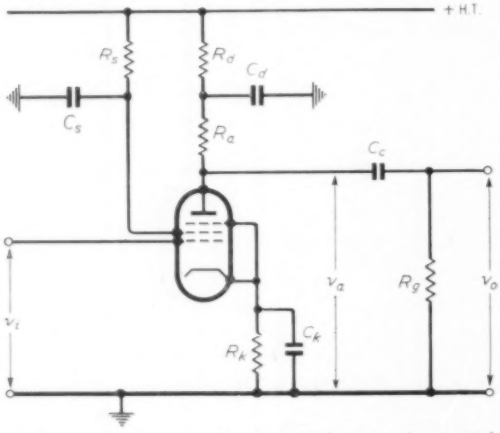

Fig. 1. *Resistance-capacitance coupled stage using pentode valve.*

3. RC Coupled Stage

Fig. 1 shows a typical resistance-capacitance coupled amplifier using a pentode valve. The usual assumptions will be made, namely:

(i) the anode and screen currents are independent of anode voltage, and the ratio betweeen them is independent of grid voltage, screen voltage and anode voltage;
(ii) the impedance of the d.c. power supply is negligible;
(iii) the grid resistance R_g is large compared with the anode load resistance R_a and decoupling resistance R_d.

Edwards and Cherry[2] have shown that the gain between grid and anode is then given by

$$\frac{v_a}{v_i} = \frac{g_m\,(R_a + Z_d)}{1 + g_k Z_k + \sigma_s Z_s} \qquad (5)$$

where g_m is the grid–anode mutual conductance.

g_k is the sum of the screen conductance and the grid–anode, grid–screen and screen–anode mutual conductances.

σ_s is the screen conductance.

Z_d, Z_k and Z_s are the impedances of the anode, cathode and screen decoupling networks respectively.

The operational expression for v_a when v_i is a unit step is therefore

$$v_a = g_m R_a \frac{1 + \dfrac{R_d}{R_a}\cdot\dfrac{1}{1+pC_dR_d}}{1 + g_k\dfrac{R_k}{1+pC_kR_k} + \sigma_s\dfrac{R_s}{1+pC_sR_s}}\,1$$

$$= G\frac{1 + \dfrac{\gamma}{p+\lambda}}{1 + g_kR_k\dfrac{\eta}{p+\eta} + \sigma_sR_s\dfrac{\theta}{p+\theta}} \qquad (6)$$

where $\gamma = 1/R_aC_d$, $\lambda = 1/R_dC_d$
$\eta = 1/R_kC_k$, $\theta = 1/R_sC_s$

and $G = g_m R_a$

$$\text{Now} \quad \frac{v_0}{v_a}1 = \frac{pC_cR_g}{1 + pC_cR_g}1 = \frac{p}{p+d}1 \qquad (7)$$

where $d = 1/R_gC_c$

It follows from equations (6) and (7) that the overall indicial response is given by

$$v_0 = G\frac{p^4 + a_1p^3 + a_2p^2 + a_3p}{p^4 + b_1p^3 + b_2p^2 + b_3p + b_4}1 \qquad (8)$$

where $a_1 = \gamma + \lambda + \eta + \theta$
$a_2 = \eta\theta + (\gamma + \lambda)(\eta + \theta)$
$a_3 = \eta\theta\,(\gamma + \lambda)$
$b_1 = \alpha + \lambda + \eta(1 + g_kR_k) + \theta\,(1 + \sigma_sR_s)$
$b_2 = \alpha\lambda + \eta(\alpha+\lambda)\,(1 + g_kR_k) + \theta\,(\alpha+\lambda)\,(1 + \sigma_sR_s) + \eta\theta\,(1 + g_kR_k + \sigma_sR_s)$
$b_3 = \alpha\lambda\eta\,(1 + g_kR_k) + \alpha\lambda\theta\,(1 + \sigma_sR_s) + \eta\theta\,(1 + g_kR_k + \sigma_sR_s)\,(\alpha+\lambda)$
$b_4 = \alpha\lambda\eta\theta\,(1 + g_kR_k + \sigma_sR_s)$

3.1. *Compensation for Anode Decoupling by means of Cathode and Screen Decoupling*

At low frequencies, the gain tends to rise because of the increase in the anode load impedance due to the finite reactance of C_d, but there is also a fall in gain due to the negative feedback introduced by the finite reactance of C_k and C_s. Edwards and Cherry[2] have shown that these effects can be made to cancel completely so that the gain between grid and anode is independent of frequency. Other circuits in which an increase in anode load impedance offsets screen and cathode feedback at low frequencies have been described by Zeidler and Noe.[3]

From equation (6) it follows that

$$v_a = G\,\mathbf{1}$$

$$\text{if} \quad \frac{\gamma}{p+\lambda} = g_k R_k \frac{\eta}{p+\eta} + \sigma_s R_s \frac{\theta}{p+\theta}$$

$$\therefore \quad \gamma = \eta\, g_k R_k + \theta \sigma_s R_s$$

$$\text{and} \quad \lambda = \eta = \theta$$

$$\text{i.e.,} \quad C_d R_d = C_k R_k = C_s R_s \quad .. \quad .. \quad (9)$$

$$\text{and} \quad R_d = R_a\,(g_k R_k + \sigma_s R_s) \quad .. \quad .. \quad (10)$$

If equations (9) are satisfied, the anode waveform is distortionless. However, unless the distortionless stage has d.c. coupling to the next stage, distortion will be introduced by the coupling capacitance. The overall indicial response is then of zero order, it being that of the coupling circuit alone, namely a simple exponential curve with time constant $1/\alpha$. This time constant must therefore be adequately large to keep the sag on the top of the waveform within the permitted limit.

3.2. *Compensation for the Coupling Circuit by means of the Anode Decoupling Circuit*

At low frequencies the gain is reduced by the increasing reactance of the coupling capacitor (corresponding to a sagging indicial response), but the increasing reactance of the anode decoupling capacitor tends to increase the gain (corresponding to a rising indicial response). The effects of the coupling and anode decoupling circuits can thus be made to offset each other. Perfect compensation cannot be obtained because the coupling circuit reduces the gain to zero at zero frequency but the increase in gain caused by the decoupling circuit is only finite. This compensating effect has been analysed by several authors.[1,4,5]

The effect of the screen and cathode decoupling can be avoided either by removing these circuits or by making their time constants long compared with those of the coupling circuit and the anode decoupling circuit. In either case (i.e., letting $R_k = R_s = 0$ or letting η and $\theta \rightarrow 0$), equation (6) becomes

$$v_a = G\left[1 + \frac{\gamma}{p+\lambda}\right]\mathbf{1} \quad .. \quad .. \quad (11)$$

Alternatively, if the screen and cathode coupling capacitors are omitted but R_s and R_k are retained, η and θ become infinite and equation (6) again reduces to the form of equation (11), but with a modified gain given by

$$G' = \frac{G}{1 + g_k R_k + \sigma_s R_s}$$

From equations (7) and (11) the overall indicial response is

$$v_0(p)\,\mathbf{1} = G\,\frac{p^2 + p\,(\gamma+\lambda)}{p^2 + p\,(\alpha+\lambda) + \alpha\lambda}\,\mathbf{1} \quad (12)$$

First-order compensation is obtained when the coefficients of p are equal in the numerator and denominator

$$\text{i.e.,} \quad \alpha = \gamma$$

$$\therefore \; R_g C_c = R_a C_d \quad .. \quad .. \quad .. \quad (13)$$

Other circuits in which an increase in anode impedance offsets the falling response of the coupling circuit at low frequencies have been described by Schlesinger.[6]

3.3. *Second-Order Compensation*

One method of obtaining second-order compensation which has been proposed by Thomson[1] involves adding further components to the coupling circuit to make it more nearly compensate for the effect of the anode decoupling circuit: the effects of the screen and cathode decoupling circuits are assumed to be negligible. It is possible, however, to obtain second-order compensation without adding any components to the circuit of Fig. 1 by making use of the screen and cathode decoupling circuits as well as the anode decoupling circuit.

From equation (8), first-order compensation is obtained when $a_1 = b_1$

$$\therefore \; \gamma = \alpha + \eta\, g_k R_k + \theta \sigma_s R_s \quad .. \quad .. \quad (14)$$

For second-order compensation it is required in addition that $a_2 = b_2$

$$\therefore \; \gamma\,(\eta+\theta) = \alpha\,(\lambda+\eta+\theta) + \eta\, g_k R_k\,(\alpha+\lambda+\theta) + \theta\sigma_s R_s\,(\alpha+\lambda+\eta) \quad .. \quad (15)$$

Eliminating γ from equation (15) by means of equation (14) gives

$$\alpha = \frac{\eta\, g_k R_k\,(\eta-\lambda) + \theta\sigma_s R_s\,(\theta-\lambda)}{\lambda + \eta\, g_k R_k + \theta\sigma_s R_s} \quad .. \quad (16)$$

If we attempt to obtain third-order compensation we also require $a_3 = b_3$ and, from equation (8) this is obtained when

$$\gamma\eta\theta = \alpha\,(\lambda\eta + \eta\theta + \theta\lambda) + \eta\, g_k R_k\,(\lambda\theta + \theta\alpha + \alpha\lambda) + \theta\sigma_s R_s\,(\lambda\eta + \eta\alpha + \alpha\lambda)$$

Eliminating γ by means of equation (14) gives

$$\eta\theta\,(\eta\,g_kR_k + \theta\sigma_sR_s) = \alpha\lambda\,[\eta\,(1 + g_kR_k) + \theta\,(1 + \sigma_sR_s)] + (\alpha + \lambda)\,\eta\theta\,(g_kR_k + \sigma_sR_s)$$

But, from equation (16),

$$\alpha + \lambda = \frac{\eta^2\,g_kR_k + \theta^2\sigma_sR_s - \alpha\lambda}{\eta\,g_kR_k + \theta\sigma_sR_s}$$

$$\therefore\ \alpha\lambda = -\frac{\eta\theta\,(\eta - \theta)^2\,g_kR_k\,\sigma_sR_s}{\eta^2\,g_kR_k\,(1 + g_kR_k) + \theta^2\sigma_sR_s(1 + \sigma_sR_s) + 2\eta\,\theta g_kR_k\,\sigma_sR_s}$$

Now α, λ, η and θ are all positive, so it is impossible to obtain third-order compensation.

The highest order of compensation which can therefore be obtained for the indicial response of the circuit shown in Fig. 1 is the second order. This is obtained when α satisfies equation (16) and γ satisfies equation (14).

3.4. *Numerical Example*

Consider, as an example, a stage using an EF91 valve. Typical parameters obtained from the published characteristics of this valve are as follows:

grid–anode mutual conductance	7·6 mA/V
grid–screen mutual conductance	2·0 mA/V
screen–anode mutual conductance	0·1 mA/V
screen conductance	0·029 mA/V

(These figures relate to the following conditions: $V_a = V_s = 250$ V, $V_g = -2$ V, $I_a = 10$ mA, $I_s = 2{\cdot}5$ mA.)

The value of g_k is therefore 9·7 mA/V. If $R_k = 160\ \Omega$ and $R_s = 18\ \text{k}\Omega$, then $g_kR_k = 1{\cdot}5$ and $\sigma_sR_s = 0{\cdot}5$.

In order to compensate for the effect of anode decoupling by means of the cathode and screen decoupling circuits, equations (9) and (10) show that the anode, cathode and screen time constants should be equal and

$$R_d = 2R_a$$

The indicial response is then of zero order, it being a simple exponential curve with the time constant of the coupling circuit.

If the effects of the screen and cathode circuits can be made negligible and the anode decoupling circuit is used to compensate for the effect of the coupling circuit ($\alpha = \gamma$), the indicial response is given by equation (12). If we keep $R_d = 2R_a$ ($\gamma = 2\lambda$) equation (12) becomes

$$v_0(p)\,1 = \frac{p^2 + 1{\cdot}5\,p\gamma}{p^2 + 1{\cdot}5\,\gamma p + 0{\cdot}5\,\gamma^2}\,1$$

whence

$$v_0(t) = 1 - 0{\cdot}25\,(\gamma t)^2 + 0{\cdot}125\,(\gamma t)^3 - 0{\cdot}0365\,(\gamma t)^4 + 0{\cdot}0078\,(\gamma t)^5 - \ldots\ldots \quad (t > 0)$$

The response calculated from this series is shown as curve B in Fig. 2. This figure also shows as curve A the response due to the coupling circuit alone (when $\alpha = \gamma$), thus showing the improvement brought about by the first-order compensation.

If all the time constants are made finite and are adjusted to give second-order compensation the indicial response is given by equation (8). If we keep $\gamma = 2\lambda$ and put $\eta = 2\theta$ then, from equations (14) and (16),

$$\alpha = 0{\cdot}016\,\gamma, \qquad \lambda = 0{\cdot}5\,\gamma$$
$$\eta = 0{\cdot}56\,\gamma, \qquad \theta = 0{\cdot}28\,\gamma$$

Equation (8) then becomes

$$v_0(p)\,1 = \frac{p^4 + 2{\cdot}34\gamma p^2 + 1{\cdot}42\gamma^2p^2 + 0{\cdot}235\gamma^3p}{p^4 + 2{\cdot}34\gamma p^3 + 1{\cdot}42\gamma^2p^2 + 0{\cdot}257\gamma^3p + 0{\cdot}00376\gamma^4}\,1$$

whence

$$v_0(t) = 1 - 0{\cdot}00037\,(\gamma t)^3 + 0{\cdot}0020\,(\gamma t)^4 - 0{\cdot}00067\,(\gamma t)^5 + \ldots\ldots$$

The response calculated from this series is shown as curve D in Fig. 2. The use of second-order compensation thus gives a great improvement over the response for first-order compensation, assuming that the value of anode time constant ($1/\gamma$) is the same in each case. However, the value of coupling time constant ($1/\alpha$) required to obtain second-order compensation is 62 times that required for first-order compensation. Fig. 2 shows as curve C the response due to the coupling circuit alone. This response is still inferior to that of curve D and it can only be obtained if the decoupling circuits are omitted, or have much larger time constants than the coupling circuit, or are arranged to compensate for each other as described in Section 3.1. If the decoupling time

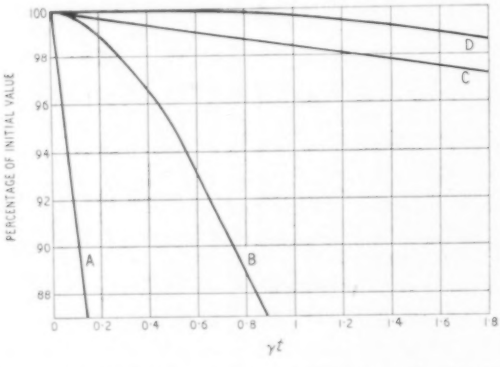

Fig. 2. *Indicial response of amplifier stage; A, due to coupling circuit alone ($\alpha = \gamma$); B, with 1st-order compensation; C, due to coupling circuit alone ($\alpha = 0{\cdot}016\gamma$); D, with 2nd-order compensation.*

constants can be chosen to compensate for each other they can be arranged to give second-order compensation as shown in curve D. The time taken by curve D to drop by 1% is over $2\frac{1}{2}$ times that taken by curve C.

4. Stages in Tandem

Let the indicial responses of two stages be

$$h_1(p)\,1 = \frac{p^n + a_1p^{n-1} + a_2p^{n-2} + \ldots + a_rp^{n-r} + \ldots + a_{n-1}p + a_n}{p^n + b_1p^{n-1} + b_2p^{n-2} + \ldots + b_rp^{n-r} + \ldots + b_{n-1}p + b_n}\,1$$

and

$$h_2(p)\,1 = \frac{p^m + a_1'p^{m-1} + a_2'p^{m-2} + \ldots + a_r'p^{m-r} + \ldots + a'_{m-1}p + a_m}{p^m + b_1'p^{m-1} + b_2'p^{m-2} + \ldots + b_r'p^{m-r} + \ldots + b'_{m-1}p + b_m}\,1$$

The overall response of the two stages in tandem is therefore

$$h(p)\,1 = h_1(p).\ h_2(p)\,1 = \frac{p^{n+m} + c_1p^{n+m-1} + \ldots + c_rp^{n+m-r} + \ldots + c_{n+m-1}p + c_{n+m}}{p^{n+m} + d_1p^{n+m-1} + \ldots + d_rp^{n+m-r} + \ldots + d_{n+m-1}p + d_{n+m}}\,1$$

where

$c_1 = a_1 + a_1'$, $\quad d_1 = b_1 + b_1'$

$c_2 = a_2 + a_1a_1' + a_2'$, $\quad d_2 = b_2 + b_1b_1' + b_2'$

$c_r = a_r + a_r' + \sum_{q=1}^{r-1} a_qa'_{r-q}$, $\quad d_r = b_r + b_r' + \sum_{q=1}^{r-1} b_qb'_{r-q}$

$c_{n+m-1} = a_{n-1}a_m' + a_na'_{m-1}$, $\quad d_{n+m-1} = b_{n-1}b_m' + b_nb'_{m-1}$

$c_{n+m} = a_na_m'$ $\quad d_{n+m} = b_nb_m'$

If one stage has rth-order compensation and the other has qth order, then

$$a_1 = b_1,\ a_2 = b_2,\ \ldots\ a_r = b_r,\ a_{r+1} \neq b_{r+1}$$

and

$$a_1' = b_1',\ a_2' = b_2',\ \ldots\ a_q' = b_q',\ a'_{q+1} \neq b'_{q+1}$$

where

$$q \geqslant r \text{ and } a_{r+1} \neq b'_{r+1},\ b_{r+1} \neq a'_{r+1}$$

therefore

$$c_1 = d_1,\ c_2 = d_2,\ \ldots\ c_r = d_r,\ c_{r+1} \neq d_{r+1}$$

The order of compensation of the complete amplifier is therefore the same as that of the stage with the lower-order compensation. It follows that if an amplifier has n stages all with rth-order compensation then the complete amplifier will also have rth-order compensation. If each stage can be given rth-order compensation, however, it is usually possible for the complete amplifier to have a higher order of compensation.

If two stages are connected in tandem and one can be made to have rth-order compensation and the other can be made to have qth-order compensation it should be possible to adjust the values of the coefficients of at least the first $(q + r)$ terms in the numerator and denominator of $h(p)$ so as to produce at least $(q + r)$th compensation. Thus it should be possible to design a multi-stage amplifier whose order of compensation is at least the sum of those which can be obtained from the individual stages. A very simple example is the tandem connection of a coupling circuit and an anode decoupling circuit. Each, considered separately, has a zero-order indicial response but when the two are connected in tandem the coefficients can be adjusted to give a first-order response as shown in Section 3.2. Some more complicated circuits are considered in Sections 4.1 to 4.4.

Because a multi-stage amplifier has a higher order of compensation than a simpler circuit it does not follow that its indicial response will approximate to the ideal more closely at all times. The multi-stage circuit with the high order of compensation will have an indicial response nearer to the ideal at times close to $t = 0$ but its response may subsequently fall more rapidly than that of the simple circuit so as to cross it and ultimately lie well below it. As an example, curve C in Fig. 2 relates to a zero-order response ($n = 1$, $r = 0$), but curve B relates to a first-order response ($n = 2$, $r = 1$). However, curve B crosses curve C quite close to $t = 0$ and, subsequently, departs from the ideal very much more rapidly.

4.1. *One Coupling Network and Two Stages with Anode Decoupling*

The operational form of the overall indicial response, from equations (7) and (11), is

$$h(p)\,1 = \frac{p}{p + \alpha}\left[1 + \frac{\gamma_1}{p + \gamma_1}\right].\left[1 + \frac{\gamma_2}{p + \gamma_2}\right]1 = \frac{p^3 + a_1p^2 + a_2p}{p^3 + b_1p^2 + b_2p + b_3}$$

where

$a_1 = \gamma_1 + \gamma_2 + \lambda_1 + \lambda_2$

$a_2 = \gamma_1\gamma_2 + \gamma_1\lambda_2 + \gamma_2\lambda_1 + \lambda_1\lambda_2$

$b_1 = \alpha + \lambda_1 + \lambda_2$

$b_2 = \alpha\lambda_1 + \alpha\lambda_2 + \lambda_1\lambda_2$

$b_3 = \alpha\gamma_1\gamma_2$

For second-order compensation we require $a_1 = b_1$ and $a_2 = b_2$

whence $\alpha = \gamma_1 + \gamma_2$ .. .. .. (17)

and $\gamma_1\gamma_2 = \gamma_1\lambda_1 + \gamma_2\lambda_2$ .. .. .. (18)

In particular, if $\gamma_1 = \gamma_2$ and $\lambda_1 = \lambda_2$ then equations (17) and (18) give

$$\alpha = 2\gamma_1 = 2\gamma_2 = 4\lambda_1 = 4\lambda_2 \quad .. \quad (19)$$

Thus, when the two anode circuits are identical, second-order compensation requires the anode decoupling resistance to be equal to twice the anode load resistance.

4.2. *Two Coupling Networks and Two Stages with Anode Decoupling*

The operational form of the overall indicial response, from equation (12) is

$$h(p)\,\mathbf{1} = \frac{p^2 + p(\gamma_1 + \lambda_1)}{p^2 + p(\alpha_1 + \lambda_1) + \alpha_1\lambda_1} \cdot \frac{p^2 + p(\gamma_2 + \lambda_2)}{p^2 + p(\alpha_2 + \lambda_2) + \alpha_2\lambda_2}\,\mathbf{1}$$
$$= \frac{p^4 + a_1p^3 + a_2p^2}{p^4 + b_1p^3 + b_2p^2 + b_3p + b_4}\,\mathbf{1}$$

where

$a_1 = \gamma_1 + \gamma_2 + \lambda_1 + \lambda_2$, $\quad b_1 = \alpha_1 + \alpha_2 + \lambda_1 + \lambda_2$

$a_2 = \gamma_1\gamma_2 + \gamma_1\lambda_2 + \gamma_2\lambda_1 + \lambda_1\lambda_2$, $\quad b_2 = \alpha_1\lambda_1 + \alpha_2\lambda_2 + (\alpha_1 + \lambda_1)(\alpha_2 + \lambda_2)$

$b_3 = \alpha_1\lambda_1(\alpha_2 + \lambda_2) + \alpha_2\lambda_2(\alpha_1 + \lambda_1)$, $\quad b_4 = \alpha_1\alpha_2\lambda_1\lambda_2$

Although the circuit considered in this section has more elements than that in Section 4.1, resulting in polynomials one degree higher for the numerator and denominator of $h(p)$, third-order compensation cannot be obtained because $a_3 = 0$.

For second-order compensation we require $a_1 = b_1$ and $a_2 = b_2$, which gives

$$\alpha_1 + \alpha_2 = \gamma_1 + \gamma_2$$

and $\quad \alpha_1\alpha_2 = \gamma_1\gamma_2 - \gamma_1\lambda_1 - \gamma_2\lambda_2$

$$\therefore\ \alpha_1, \alpha_2 = \tfrac{1}{2}\{(\gamma_1 + \gamma_2) \pm \sqrt{[(\gamma_1 - \gamma_2)^2 + 4(\gamma_1\lambda_1 + \gamma_2\lambda_2)]}\} \quad .. \quad .. \quad (20)$$

Equation (20) cannot have equal roots, so second-order compensation cannot be obtained when the coupling circuits have equal time constants. The anode circuits can be identical however. In this particular case $\gamma_1 = \gamma_2 = \gamma$ and $\lambda_1 = \lambda_2 = \lambda$. Equation (20) then gives

$$\alpha_1, \alpha_2 = \gamma \pm \sqrt{2\gamma\lambda} \quad .. \quad .. \quad (21)$$

Second-order compensation can therefore be obtained provided that $\gamma > 2\lambda$. The anode decoupling resistance can thus have any value greater than $2R_a$.

4.3. *m Coupling Networks and n Stages with Anode Decoupling*

The operational form of the indicial response from equations (7) and (11) is

$$h(p)\,\mathbf{1} = \prod_{r=1}^{m} \frac{p}{p + \alpha_r} \cdot \prod_{s=1}^{n} \frac{p + \gamma_s + \lambda_s}{p + \lambda_s}\,\mathbf{1} = \frac{p^{m+n} + a_1p^{m+n-1} + \ldots + a_np^m}{p^{m+n} + b_1p^{m+n-1} + \ldots + b_np^m + \ldots + b_{m+n}}\,\mathbf{1}$$

This shows that the maximum order of compensation that can be achieved is n, the number of decoupling circuits. Coupling circuits do not increase the order of compensation, although they may assist in providing a sufficient number of variables to make a desired order of compensation realizable.

In order to obtain a response with a higher order than zero, there must be at least one coupling circuit. If there is not (i.e., if $m = 0$), then

$$a_1 = \sum_{s=1}^{n} (\gamma_s + \lambda_s), \qquad b_1 = \sum_{s=1}^{n} \lambda_s$$

The condition $a_1 = b_1$ gives $\sum_{s=1}^{n} \gamma_s = 0$, which is impossible because all γ are positive.

It can be shown that one coupling circuit is also sufficient to obtain nth-order compensation for at least $n = 1, 2, 3, 4, 5$. The cases $n = 1$ and $n = 2$ have been dealt with in Sections 2.2 and 4.1. A further example, $n = 3$, is given in Section 4.4.

4.4. *One Coupling Network and Three Stages with Anode Decoupling*

The operational form of the indicial response, from equations (7) and (11) is

$$h(p)\,\mathbf{1} = \frac{p}{p + \alpha} \cdot \frac{p + \gamma_1 + \lambda_1}{p + \lambda_1} \cdot \frac{p + \gamma_2 + \lambda_2}{p + \lambda_2} \times \frac{p + \gamma_3 + \lambda_3}{p + \lambda_3}$$
$$= \frac{p^4 + a_1p^3 + a_2p^2 + a_3p}{p^4 + b_1p^3 + b_2p^2 + b_3p + b_4}\,\mathbf{1}$$

where

$a_1 = \gamma_1 + \gamma_2 + \gamma_3 + \lambda_1 + \lambda_2 + \lambda_3$

$a_2 = (\gamma_1 + \lambda_1)(\gamma_2 + \lambda_2) + (\gamma_2 + \lambda_2)(\gamma_3 + \lambda_3) + (\gamma_3 + \lambda_3)(\gamma_1 + \lambda_1)$

$a_3 = (\gamma_1 + \lambda_1)(\gamma_2 + \lambda_2)(\gamma_3 + \lambda_3)$

$b_1 = \alpha + \lambda_1 + \lambda_2 + \lambda_3$

$$b_2 = \lambda_1\lambda_2 + \lambda_2\lambda_3 + \lambda_3\lambda_1 + \alpha(\lambda_1 + \lambda_2 + \lambda_3)$$
$$b_3 = \lambda_1\lambda_2\lambda_3 + \alpha(\lambda_1\lambda_2 + \lambda_2\lambda_3 + \lambda_3\lambda_1)$$
$$b_4 = \alpha\lambda_1\lambda_2\lambda_3$$

For first-order compensation $a_1 = b_1$

$$\therefore \quad \alpha = \gamma_1 + \gamma_2 + \gamma_3 \quad .. \quad .. \quad .. \quad (22)$$

For second-order compensation, putting $a_2 = b_2$ and eliminating α by means of equation (22) gives

$$\gamma_1\gamma_2 + \gamma_2\gamma_3 + \gamma_3\gamma_1 = \gamma_1\lambda_1 + \gamma_2\lambda_2 + \gamma_3\lambda_3 \quad (23)$$

Third-order compensation requires in addition $a_3 = b_3$ which, after eliminating α by means of equation (22) gives

$$\gamma_1\lambda_1(\lambda_2 + \lambda_3) + \gamma_2\lambda_2(\lambda_1 + \lambda_3) + \gamma_3\lambda_3(\lambda_1 + \lambda_2)$$
$$= \lambda_1\gamma_2\gamma_3 + \lambda_2\gamma_1\gamma_3 + \lambda_3\gamma_1\gamma_2 + \gamma_1\gamma_2\gamma_3 \quad .. \quad (24)$$

Third-order compensation is, therefore, obtained when equations (23) and (24) are satisfied. It can be shown that these equations cannot be satisfied when $\lambda_1 = \lambda_2 = \lambda_3$ but they can be satisfied when $\gamma_1 = \gamma_2 = \gamma_3$. In the particular case when $\gamma_1 = \gamma_2 = \gamma_3 = \gamma$ and $\lambda_1 = \lambda_2 = \lambda$, equations (23) and (24) become

$$3\gamma = 2\lambda + \lambda_3$$

and $$\gamma^2 + \gamma(2\lambda + \lambda_3) = 2\lambda^2 + 4\lambda\lambda_3$$

which give

$$\gamma_1 = \gamma_2 = \gamma_3 = \gamma = \lambda_3(2\sqrt{3} - 3) \quad .. \quad (25)$$
$$\lambda_1 = \lambda_2 = \lambda = \lambda_3(3\sqrt{3} - 5) \quad .. \quad (26)$$
$$\alpha = \lambda_3(6\sqrt{3} - 9) \quad .. \quad (27)$$

The values of the coefficients are given in the appendix.

The gain–frequency response will be said to have rth-order compensation when the first r derivatives of $|g(\omega)|$ with respect to ω vanish at infinite frequency. Similarly, the phase–frequency response will be said to have rth-order compensation when the first r derivatives of $\phi(\omega)$ vanish at infinite frequency. The gain–frequency response will be called maximally-flat when it has $(2n-1)$th-order compensation and the phase–frequency response will be called maximally flat when it has $(2n-2)$th-order compensation.

It is shown in the appendix that when the indicial response has compensation of order $2r$ or $(2r + 1)$ the gain–frequency response has compensation of order $(2r + 1)$. The maximally-flat indicial response, therefore, always corresponds to a gain–frequency response with a lower order of compensation than that required for maximal flatness.

When the indicial response has compensation of order $2r$ or $(2r - 1)$ the phase response has compensation of order $2r$. Thus only when $n = 2$ does the maximally-flat indicial response correspond to a maximally-flat phase response. For all higher values of n the maximally-flat indicial response corresponds to a phase response of a lower order of compensation than that required for maximal flatness. This is illustrated by the examples shown in Table 1.

TABLE 1

Order of max. flat indicial response	Order of corresponding gain-freq. response	Order of corresponding phase-freq. response	Order of max. flat gain-freq. response	Order of max. flat phase-freq. response
1	1	2	3	2
2	3	2	5	4
3	3	4	7	6
4	5	4	9	8

5. Comparison with Gain and Phase Responses

It is shown in Appendix 2 that the gain–frequency response corresponding to the indicial response given by equation (1) is

$$|g(\omega)| = \left[\frac{\omega^{2n} + c_2\omega^{2n-2} + \ldots + c_{2r}\omega^{2n-2r} + \ldots + c_{2n}}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2n-2r} + \ldots + d_{2n}}\right]^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad .. \quad (28)$$

and the phase–frequency response is given by

$$\phi(\omega) = \tan^{-1}\left[\frac{c_1\omega^{2n-1} + c_3\omega^{2n-3} + \ldots + c_{2n-1}\omega}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2n-2r} + \ldots + d_{2n}}\right] \quad .. \quad .. \quad .. \quad .. \quad (29)$$

Consider, as an example, the indicial response given by

$$h(p)\,\mathbf{1} = \frac{p^3 + a_1p^2 + a_2p}{p^3 + b_1p^2 + b_2p + b_3}\,\mathbf{1}$$

This has a maximally-flat indicial response, which is second-order, when $a_1 = b_1$ and $a_2 = b_2$.

The corresponding gain–frequency response, from equation (28), is given by

$$|g(\omega)|^2 = \frac{\omega^6 + c_2\omega^4 + c_4\omega^2}{\omega^6 + d_2\omega^4 + d_4\omega^2 + d_6}$$

where the coefficients, given by Appendix 2.1 are

$$c_2 = a_1^2 - 2a_2\,, \qquad d_2 = b_1^2 - 2b_2$$
$$c_4 = a_2^2\,, \qquad d_4 = b_2^2 - 2b_3b_1$$
$$d_6 = b_3^2$$

This has a maximally-flat gain–frequency response, which is fifth-order, when $c_2 = d_2$ and $c_4 = d_4$.

$$\therefore \quad b_1 = \sqrt{[a_1^2 + 2(a_2 - b_2)]}$$

and $$b_3 = [a_2^2 - b_2^2]/2\sqrt{[a_1^2 + 2(a_2 - b_2)]}$$

The phase–frequency response, from equation (29), is given by

$$\tan \phi(\omega) = \frac{c_1\omega^5 + c_3\omega^3 + c_5\omega}{\omega^6 + d_2\omega^4 + d_4\omega^2}$$

where the coefficients, given by Appendix 2.2, are

$$c_1 = b_1 - a_1 \qquad d_2 = a_1b_1 - a_2 - b_2$$
$$c_3 = a_1b_2 - a_2b_1 - b_3 \qquad d_4 = a_2b_2 - a_1b_3$$
$$c_5 = b_3a_2$$

This has a maximally-flat phase–frequency response, which is fourth-order, when

$$c_1 = c_3 = 0$$
$$\therefore b_1 = a_1$$

and $$b_2 = a_2 + b_3/a_1$$

Thus the condition for a maximally-flat indicial response corresponds neither to the maximally-flat gain–frequency response nor to the maximally-flat phase–frequency response. The latter, however, does give a first-order compensated indicial response whereas the maximally-flat gain–frequency response corresponds to an uncompensated (i.e., zero-order) indicial response.

6. Experimental Work

Fig. 3 shows an amplifier which was built in order to obtain some experimental verification of the theory. The values of the components determining the time-constants of the coupling circuits and decoupling circuits are accurate to $\pm 1\%$. The time-constants of the screen decoupling circuits are of the order of 100 times those of the coupling circuits and anode decoupling circuits, which were made much smaller than normal in order to avoid distortion being introduced by the oscilloscope amplifier. The effects of time-constants other than those of the two coupling circuits and two anode decoupling circuits are, therefore, negligible and the analysis given in Section 4.2 is applicable. The voltage divider before each stage reduces its gain to about unity in order to facilitate the taking of oscillograms.

Oscillograms were taken of the response of the amplifier to a 50-c/s square-wave input voltage. Because the period of this wave is long compared with the amplifier time-constants, the response to each edge of the waveform is substantially independent of its predecessors and approximates closely to the indicial response. A double-beam oscilloscope was used with the output voltage deflecting one beam and a 500-c/s sine-wave calibrating signal deflecting the other beam.

When each coupling capacitor is $0{\cdot}0125\,\mu$F the time-constants of the coupling and decoupling circuits are equal. Each stage should then have first-order compensation of its indicial response, and so should the complete amplifier. Figs. 4(a) and (b) show the response of each stage separately and Fig. 4(c) shows the response of the two stages in tandem. The initial flatness of the response indicates that first-order compensation was obtained in each case.

When the coupling capacitances are $0{\cdot}00688\,\mu$F and $0{\cdot}0682\,\mu$F respectively, the time-constants have the values given by equation (21). The complete amplifier should then have second-order compensation. Figs. 4 (d) and (e) show the response of each stage and (f) shows the response of the two stages in tandem. The response in Fig. 4(f) maintains its initial flatness longer than that in Fig. 4(c), as is expected with second-order

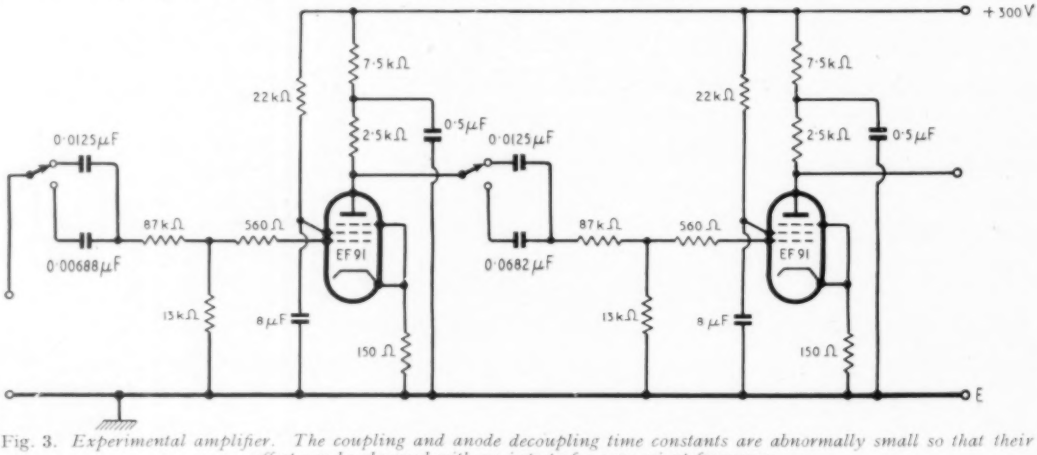

Fig. 3. *Experimental amplifier. The coupling and anode decoupling time constants are abnormally small so that their effect can be observed with an input of a convenient frequency.*

compensation. The total amount by which the top of the response sags in 2 ms in the second-order case is about one-third of that with first-order compensation.

The time-constants obtained in practice will depart from the calculated values because of the tolerances on components. Fig. 5 shows the effect on the overall response of a $\pm 10\%$ variation of the coupling capacitance of the first stage. The response of the amplifier with incorrectly-adjusted second-order compensation compares favourably with that obtained with correctly-adjusted first-order compensation.

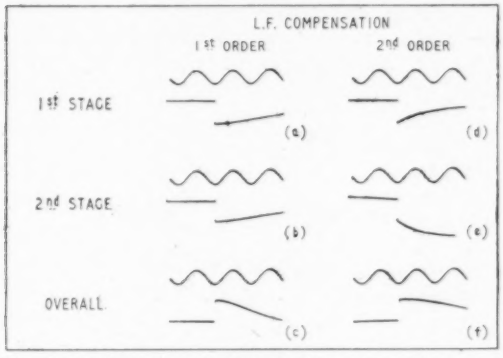

Fig. 4. *Oscillograms of response of model amplifier to a 50-c/s square wave. (The timing wave is 500 c/s.)*

The gain–frequency response of the amplifier was measured with the component values corresponding to first- and second-order indicial responses. The results obtained are shown in Fig. 6. In each case the response is far from being maximally flat, thus confirming that a flat gain–frequency response is not the criterion for a flat indicial response. The frequency at which the second-order response has fallen 3 db is nearly half of that for the first-order case but the peak in the second-order curve is higher.

7. Conclusion

The indicial response of an amplifier can be made maximally flat by causing as many as possible of its differential coefficients to be zero at $t = 0$. A general method for doing this has been developed and applied to some typical circuits.

Some experimental verification has also been obtained. In order to obtain as high an order of compensation as possible, in many cases, it is necessary for the anode decoupling resistance to be several times the anode load resistance. This is quite practicable in wideband amplifiers without the use of an excessive supply voltage because the anode load resistance must be small in order to obtain adequate high-frequency performance.

A single resistance-capacitance coupled stage can have second-order compensation. As the number of stages in an amplifier is increased, the order of compensation which can be obtained increases, but so does the complexity of the calculations required to obtain the conditions for compensation. In many practical cases it is sufficient to obtain first- or second-order compensation. A complicated circuit can then be dealt with by considering it as a number of simpler stages connected in tandem, each of which can be made to have the required order of compensation.

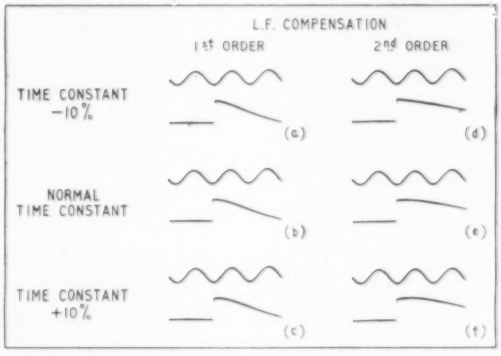

Fig. 5. *Effect of varying the time constant of one coupling circuit on the response of model amplifier to a 50-c/s square wave. (The timing wave is 500 c/s.)*

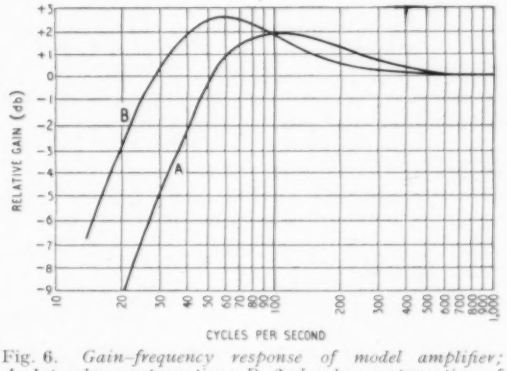

Fig. 6. *Gain–frequency response of model amplifier; A, 1st-order compensation; B, 2nd-order compensation of indicial response.*

APPENDIX 1

The substitution $x = 1/p$ in equation (1) gives

$$h(x) = \frac{1 + a_1 x + a_2 x^2 + \ldots + a_r x^r + \ldots + a_n x^n}{1 + b_1 x + b_2 x^2 + \ldots + b_r x^r + \ldots + b_n x^n} = \frac{1 + A}{1 + B}$$

where $A = \sum_{r=1}^{n} a_r x^r$ and $B = \sum_{r=1}^{n} b_r x^r$

$$\therefore h(x) = (1 + A)[1 - B + B^2 - \ldots + (-1)^r B^r + \ldots] \quad (30)$$

$$= 1 + c_1 x + c_2 x^2 + \ldots + c_r x^r + \ldots$$

Let $[1 - B + \ldots + (-1)^r B^r \ldots] = 1 + d_1 x + d_2 x^2 + \ldots + d_r x^r + \ldots$ then by the multinomial theorem

$$d_r = \sum \frac{(-1)^m m!}{\alpha!\,\beta! \ldots \epsilon!} b_\lambda{}^\alpha \cdot b_\mu{}^\beta \ldots \ldots b_\rho{}^\epsilon \quad (4)$$

where the summation extends over all possible values of $\alpha, \beta, \ldots \epsilon$ and $\lambda, \mu \ldots \rho$ and where

$m = \alpha + \beta + \ldots + \epsilon$; $\alpha\lambda + \beta\mu + \ldots + \epsilon\rho = r$; $1 \leqslant \lambda < \mu < \ldots < \rho \leqslant r$

Then from equation (30)

$$c_r = d_r + \sum_{s=0}^{r-1} d_s a_{r-s}$$

It is proved below that

$$d_r = -\sum_{s=0}^{r-1} d_s b_{r-s} \quad (31)$$

it therefore follows that

$$c_r = \sum_{s=0}^{r-1} d_s a_{r-s} - \sum_{s=0}^{r-1} d_s b_{r-s}$$

$$= \sum_{s=0}^{r-1} d_s (a_{r-s} - b_{r-s}) \quad (3)$$

where $a_0 = b_0 = d_0 = 1$

To prove equation (31), assume it to be true for a particular value of r and derive from it the value of d_{r+1}. In each factor $b_\lambda{}^\alpha \cdot b_\mu{}^\beta \ldots b_\rho{}^\epsilon$ (where $\alpha\lambda + \beta\mu + \ldots + \epsilon\rho = r$ and $\alpha + \beta + \ldots + \epsilon = m$) one of the m factors b_q is replaced by b_{q+1}, so that a term like $d_s b_{r-s}$ is replaced by $d_s b_{r+1-s}$; in addition, each term in d_r occurs multiplied by b_1. Also, when like terms are collected, the numerical factor of $b_\lambda{}^{\alpha'} \cdot b_\mu{}^{\beta'} \ldots b_\rho{}^{\epsilon'}$ (where $\alpha' + \beta' + \ldots + \epsilon' = m + 1$ and $\alpha'\lambda' + \beta'\mu' + \ldots + \epsilon'\rho' = r + 1$) is

$$\frac{m!}{(\alpha' - 1)!\beta'! \ldots \epsilon'!} + \frac{m!}{\alpha'!(\beta' - 1)! \ldots \epsilon'!} + \ldots + \frac{m!}{\alpha'!\,\beta'! \ldots (\epsilon' - 1)!}$$

$$= \frac{m!}{\alpha'!\,\beta'! \ldots \epsilon'!}[\alpha' + \beta' + \ldots + \epsilon']$$

$$= \frac{m!(m+1)}{\alpha'!\,\beta'! \ldots \epsilon'!} = \frac{(m+1)!}{\alpha'!\,\beta'! \ldots \epsilon'!}$$

which is the correct multinomial coefficient, so that

$$d_{r+1} = -\left\{\sum_{s=0}^{r-1} d_s b_{r+1-s} + d_r b_1\right\} = -\sum_{s=0}^{r} d_s b_{r+1-s}$$

which is of the same form as d_r [given by equation (31)] with r replaced by $r + 1$ throughout. The values of d_r for $r = 1, 2, 3$ give the values of c_1, c_2 and c_3 shown in Section 2. These values are easily obtained from equation (1) by direct division, thus verifying equations (3) and (4) for $r = 1$, 2 and 3. But it has been shown that if equation (31) holds for r it holds for $r + 1$. Therefore, by induction, equations (3) and (4) are valid for all positive integral values of r.

APPENDIX 2

The steady-state complex gain, from equation (1) is

$$g(\omega) \angle \phi(\omega) = G\frac{a_0(j\omega)^n + a_1(j\omega)^{n-1} + \ldots + a_r(j\omega)^{n-r} + \ldots + a_n}{b_0(j\omega)^n + b_1(j\omega)^{n-1} + \ldots + b_r(j\omega)^{n-r} + \ldots + b_n}$$

where $a_0 = b_0 = 1$

$$= G\frac{A + jB}{C + jD} \quad (32)$$

If n is even ($n = 2m$) then

$$A = \sum_{r=0}^{m} a_{2r}(-\omega^2)^{m-r} \qquad B = \omega\sum_{r=1}^{m} a_{2r-1}(-\omega^2)^{m-r}$$

$$C = \sum_{r=0}^{m} b_{2r}(-\omega^2)^{m-r} \qquad D = \omega\sum_{r=1}^{m} b_{2r-1}(-\omega^2)^{m-r}$$

If n is odd ($n = 2m + 1$) then

$$A = \sum_{r=0}^{m} a_{2r+1}(-\omega^2)^{m-r} \qquad B = \omega\sum_{r=0}^{m} a_{2r}(-\omega^2)^{m-r}$$

$$C = \sum_{r=0}^{m} b_{2r+1}(-\omega^2)^{m-r} \qquad D = \omega\sum_{r=0}^{m} b_{2r}(-\omega^2)^{m-r}$$

2.1. *Gain–Frequency Response*

From equation (32)

$$|g(\omega)|^2 = \frac{A^2 + B^2}{C^2 + D^2}$$

$$= \frac{\omega^{2n} + c_2\omega^{2n-2} + \ldots + c_{2r}\omega^{2n-2r} + \ldots + c_{2n}}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2n-2r} + \ldots + d_{2n}} \quad (28)$$

If n is even ($n = 2m$)

$$A^2 = \sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} a_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ (where $r = q + s$) is

$$(-1)^r \sum_{q=0}^{r} a_{2q}a_{2(r-q)}$$

$$B^2 = \omega^2 \sum_{q=1}^{m} a_{2q-1}(-\omega^2)^{m-q} \cdot \sum_{s=1}^{m} a_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ is

$$(-1)^{r+1} \sum_{q=1}^{r} a_{2q-1}a_{2(r-q)+1}$$

$$\therefore A^2 + B^2 = \omega^{2n} + \sum_{r=1}^{n} c_{2r}\omega^{2(n-r)}$$

where $c_{2r} = (-1)^r \sum_{q=0}^{r} (a_{2q}a_{2(r-q)} - a_{2q-1}a_{2(r-q)+1})$

$$= (-1)^r \sum_{q=0}^{2r} (-1)^q a_q a_{2r-q}$$

$$= 2\sum_{s=0}^{r} (-1)^s a_{r+s}a_{r-s} \quad (34)$$

Similarly, $C^2 + D^2 = \omega^{2n} + \sum_{r=1}^{n} d_{2r}\omega^{2(n-r)}$

where $$d_{2r} = 2\sum_{s=0}^{r}(-1)^s b_{r+s}b_{r-s} \quad .. \quad .. \quad (35)$$

If n is odd $(n = 2m + 1)$

$$A^2 = \sum_{q=0}^{m} a_{2q+1}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} a_{2s+1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ is

$$(-1)^{r+1}\sum_{q=0}^{r-1} a_{2q+1}a_{2(r-q)-1}$$

$$B^2 = \omega^2\sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} a_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ is

$$(-1)^r\sum_{q=0}^{r} a_{2q}a_{2(r-q)}$$

$$\therefore\ c_{2r} = (-1)^r\sum_{q=0}^{r}(a_{2q}a_{2(r-q)} - a_{2q-1}a_{2(r-q)+1})$$

$$= 2\sum_{s=0}^{r}(-1)^s a_{r+s}a_{r-s}$$

and similarly for d_{2q}.

The coefficients c_{2q} and d_{2q} therefore have the same values whether n is even or odd.

Putting $1/\omega = x$ in equation (28) and expanding it ascending powers of x gives

$$|g(\omega)|^2 = 1 + e_2x^2 + e_4x^4 + \ldots + e_{2r}x_2{}^r + \ldots$$

where $e_{2r} = \sum_{s=0}^{r-1} g_s(c_{2(r-s)} - d_{2(r-s)})$ .. .. .. (36)

and $g_s = \sum(-1)^m \frac{m!}{\alpha!\,\beta!\ldots\epsilon!} d_{2\lambda}{}^{\alpha}d_{2\mu}{}^{\beta}\ldots d_{2\rho}{}^{\epsilon}$ (37)

where the sum extends over all possible values of α, β $\ldots$ ϵ and λ, μ $\ldots$ ρ and

$$\alpha + \beta + \ldots + \epsilon = m\ ,\ \alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s$$

Now by Maclaurin's theorem $e_{2r} = \frac{1}{(2r)!}f_{2r}(0)$

and $$f_{2r+1}(0) = 0$$

where $f_r(0)$ is the rth differential coefficient of $|g(\omega)^2|$ at $x = 0$ (i.e., $\omega = \infty$). The gain–frequency response therefore has $(2r + 1)$th-order compensation when $e_2 = e_4 \ldots e_{2r} = 0$. Equation (36) shows that this occurs when $c_1 = d_1$, $c_2 = d_2 \ldots c_{2r} = d_{2r}$.

If the indicial response has compensation of order $2r$ or $(2r + 1)$ then $a_q = b_q$ for $q = 1, 2 \ldots 2r$ or $q = 1, 2 \ldots (2_r + 1)$ respectively. It follows from equations (34) and (35) that $c_1 = d_1$, $c_2 = d_2, \ldots c_{2r} = d_{2r}$. Therefore, from equation (36), $e_2 = e_4 = \ldots = e_{2r} = 0$. Thus, when the indicial response has compensation of order $2r$ or $(2r + 1)$ the gain–frequency response has compensation of order $(2r + 1)$. The converse is not true, because the condition $e_2 = e_4 = \ldots = e_{2r}$ is insufficient to determine the coefficients $a_1 \ldots a_r$ and $b_1 \ldots b_r$ uniquely.

2.2. *Phase–Frequency Response*

From equation (32)

$$\tan\phi(\omega) = \frac{BC - AD}{AC + BD} \quad .. \quad .. \quad .. \quad .. \quad (38)$$

$$= \frac{c_1\omega^{2n-1} + c_3\omega^{2n-3} + \ldots + c_{2r-1}\omega^{2(n-r)+1} + \ldots + c_{2n-1}\omega}{\omega^{2n} + d_2\omega^{2n-2} + \ldots + d_{2r}\omega^{2(n-r)} + \ldots + d_{2n}} \quad .. \quad .. \quad (29)$$

If n is even $(n = 2m)$

$$BC = \omega\sum_{q=0}^{m} b_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=1}^{m} a_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ (where $r = q + s$) is

$$(-1)^r\sum_{q=1}^{r} a_{2r-1}b_{2(r-q)}$$

$$AD = \omega\sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=1}^{m} b_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ is

$$(-1)^r\sum_{q=1}^{r} a_{2(r-q)}b_{2q-1}$$

$$\therefore\ BC - AD = \sum_{r=1}^{n} c_{2r-1}\omega^{2(n-r)+1}$$

where

$$c_{2r-1} = (-1)^r\sum_{q=1}^{r}(a_{2q-1}b_{2(r-q)} - a_{2(r-q)}b_{2q-1}) \quad .. \quad (39)$$

$$AC = \sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} b_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in AC is

$$(-1)^r\sum_{q=0}^{r} a_{2q}b_{2(r-q)}$$

$$BD = \omega^2\sum_{q=1}^{m} a_{2q-1}(-\omega^2)^{m-q} \cdot \sum_{s=1}^{m} b_{2s-1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in BD is

$$(-1)^{r+1}\sum_{q=1}^{r} a_{2q-1}b_{2(r-q)+1}$$

$$\therefore\ AC + BD = \sum_{r=0}^{n} d_{2r}\omega^{2(n-r)}$$

where $d_0 = 1$.

and

$$d_{2r} = (-1)^r\left\{a_0b_{2r} + \sum_{q=1}^{r}(a_{2q}b_{2(r-q)} - a_{2q-1}b_{2(r-q)+1})\right\}$$

.. .. .. (40)

If n is odd $(n = 2m + 1)$

$$BC = \omega\sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q} \cdot \sum_{q=0}^{m} b_{2s+1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ in BC is

$$(-1)^{r-1}\sum_{q=0}^{r-1} b_{2(r-q)-1}a_{2q}$$

$$AD = \omega\sum_{q=0}^{m} a_{2q+1}(-\omega^2)^{m-q} \cdot \sum_{s=0}^{m} b_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)+1}$ in AD is

$$(-1)^{r-1}\sum_{q=0}^{r-1} a_{2(r-q)-1}b_{2q}$$

$$\therefore\ c_{2r-1} = (-1)^{r-1}\sum_{q=0}^{r-1}(a_{2q}b_{2(r-q)-1} - a_{2(r-q)-1}b_{2q})$$

$$= (-1)^{r}\sum_{q=1}^{r}(a_{2q-1}b_{2(r-q)} - a_{2(r-q)}b_{2q-1})$$

$$AC = \sum_{q=0}^{m} a_{2q+1}(-\omega^2)^{m-q}\sum_{s=0}^{m} b_{2s+1}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in AD is

$$(-1)^{r-1}\sum_{q=0}^{r-1} a_{2q+1}b_{2(r-q)-1}$$

$$BD = \omega^2\sum_{q=0}^{m} a_{2q}(-\omega^2)^{m-q}\ .\ \sum_{s=0}^{m} b_{2s}(-\omega^2)^{m-s}$$

The coefficient of $\omega^{2(n-r)}$ in BD is

$$(-1)^{r}\sum_{q=0}^{r} a_{2q}b_{2(r-q)}$$

$$\therefore\ d_{2r} = (-1)^{r}\left\{a_0b_{2r} + \sum_{q=1}^{r}(a_{2q}b_{2(r-q)} - a_{2q-1}b_{2(r-q)+1})\right\}$$

The coefficients c_{2r-1} and d_{2r} therefore have the same values whether n is even or odd.

Putting $y = \omega^2$ in equation (29)

$$\tan\phi = \frac{c_1}{\omega}\cdot\frac{y^m + \frac{c_3}{c_1}y^{m-1} + \frac{c_5}{c_1}y^{m-2} + \ldots + \frac{c_{2r+1}}{c_1}y^{m-r} + \ldots + \frac{c_{2n-1}}{c_1}y}{y^m + d_2y^{m-1} + d_4y^{m-1} + \ldots + d_{2r}y^{m-r} + \ldots + d_{2n}}$$

Expanding this in ascending powers of $1/y$ gives

$$\tan\phi = \frac{c_1}{\omega}\left(1 + \frac{e_1}{y} + \frac{e_2}{y^2} + \ldots + \frac{e_r}{y^r} + \ldots\right)$$

$$= k_1x + k_3x^3 + \ldots + k_{2r+1}x^{2m+1} + \ldots$$

where $x = 1/\omega$

and $\quad k_{2r+1} = c_1e_r = \sum_{s=0}^{r-1} g_s(c_{2r+1-s} - c_1d_{2r-s}) \quad \ldots \quad (41)$

and $\quad g_s = \sum(-1)^m\frac{m!}{\alpha!\ \beta!\ \ldots\ \epsilon!}d_{2\lambda}{}^{\alpha}d_{2\mu}{}^{\beta}\ \ldots\ d_{2\rho}{}^{\epsilon} \quad (42)$

where the sum extends over all possible values of α, $\beta\ \ldots\ \epsilon$ and $\alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s$

and $\alpha + \beta + \ldots + \epsilon = m$; $\alpha\lambda + \beta\mu + \ldots + \epsilon\rho = s$

Now by Maclaurin's theorem

$$k_{2r+1} = \frac{1}{(2r+1)!}f_{2r+1}(0)$$

and $\quad f_{2r}(0) = 0$

where f_r is the rth differential coefficient of $\tan\phi$ at $x = 0(\omega = \infty)$

The phase–frequency response therefore has $2r$th-order compensation when $k_1 = k_3 = \ldots = k_{2r-1} = 0$. Equation (41) shows that this occurs when $c_1 = c_3 = \ldots = c_{2r-1} = 0$.

If the indicial response has compensation of order $(2r-1)$ or $2r$ then $a_q = b_q$ for $q = 1, 2\ \ldots\ (2r-1)$ or $q = 1, 2, \ldots\ 2r$ respectively. It follows from equation (39) that $c_1 = c_3 = \ldots = c_{2r-1} = 0$. Thus, when the indicial response has compensation of order $(2r-1)$ or $2r$ the phase–frequency response has compensation of order $(2r-1)$. The converse is not true since the condition $c_1 = c_3 = \ldots c_{2r-1} = 0$ is insufficient to determine the coefficients $a_1\ldots a_r$ and $b_1\ldots b_r$ uniquely.

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AUTOMATIC TUNING FOR PRIMARY RADAR

By S. Ratcliffe, B.Sc.

(*Concluded from p. 165, June issue*)

Choice of A.F.C. System

In drawing up the requirements for the a.f.c. system, it is important to avoid placing an unnecessary load on the a.f.c. by calling on it to compensate for a defect elsewhere in the equipment which may be more readily dealt with by attention to the design of the component in question. For example, a thermostat in series with the blower motor may be used to reduce temperature drifts in the magnetron. The a.f.c. designer should, therefore, avoid being tempted to display his ingenuity in solving problems of this type, since the increased complexity of the a.f.c. usually brings with it an increased risk of component failures. Alternatively, it is necessary to avoid lowering the reliability of the a.f.c. by attempting to perform a complex task with too few valves and too many preset adjustments.

It is necessary, in particular, to decide whether or not to make any serious attempt to deal with scanner pulling. In general, the wider the pulse and the higher the transmitter power, the more likely it is that the improvement in performance will justify the increased complexity. It is then also possible to decide whether to use a narrow (i.e., electronic) or wide-band a.f.c. system. The objections to a narrow-range system have been given earlier, and in an airborne equipment the arguments in favour of a wide-range system are very strong indeed. On a ground or shipborne equipment where a mechanic is in attendance, there is more difference of opinion. The argument may be summarized by saying that a human being usually shows more signs of intelligence but often also of fatigue.

The choice of local oscillator will be influenced by many factors which have little connection with a.f.c., but care should be taken to avoid the trap of seeking an oscillator with the widest possible tuning range, regardless of its other demerits. The most important thing, from the point of view of the a.f.c. design, is the difference between the available electronic-tuning range and the possible drift in local-oscillator frequency due to changes in temperature and pressure. If a wide-range reflector-cum-mechanical a.f.c. is to be used, it is advisable to avoid flexible mechanical couplings between the tuning mechanism and the control motor and gear train. This is best achieved by making the resonant cavity an integral part of the radar plumbing, with a plug-in klystron. Valves designed for use in this way are now becoming available.

The typical circuits given previously, together with the theoretical analysis given in the Appendix, should assist in the choice between the 'instantaneous' and the 'extrapolating' types of reflector a.f.c. If a pure reflector a.f.c. is being designed, it may be necessary to provide a search valve, either to get an adequate pull-in range, or to enable the circuit to find a suitable reflector voltage to obtain oscillations. Two main types of search circuit have been described, and the ingenious reader may readily devise many variants on either of these themes. The essential point is to arrive at a designable circuit in which the bare minimum of preset controls is required.

It must again be pointed out that it is essential to take every possible step to keep down tolerances on the gain round the reflector a.f.c. loop. Even if the transient response of the system is unimportant, the problems set by the harmonic output of the mixer will usually necessitate a close control of the gain, up to the search-stopping mechanism at least. This point must be clearly understood. It is completely erroneous to argue from normal negative-feedback principles, but it is a common misconception that gain is not critical above a useful level.

Performance Tests on an A.F.C. Loop

Assuming that a design has been completed and an experimental model is available, it is necessary to check that the performance is satisfactory.

Consider first the problem of determining the locking accuracy. A very convenient means of checking this involves the use of a spectrum analyser of the type consisting of a narrow-band superheterodyne receiver having a local oscillator which is frequency modulated by means of a low-frequency sawtooth, and a c.r.t. display on which the receiver response is shown as a function of frequency. The intermediate frequency of this receiver is made exactly half the i.f. of the a.f.c. under test, and the spectrum analyser is supplied with a signal input both from the magnetron of the radar, and from the local oscillator to which the a.f.c. is being applied. Such a spectrum analyser will, at any given point in its sweep, respond to signals on two frequencies

one above and the other below that of the analyser oscillator, and the difference of the two frequencies will be twice the i.f. of the spectrum analyser; i.e., equal to the radar receiver i.f. If, then, the a.f.c. under test is holding a local oscillator at the correct frequency separation from a pulse source, it is possible, with an analyser such as that described, to display the spectra of the pulse and of the local-oscillator output superimposed on each other. Fig. 24 should help to make this clear. Any failure of the a.f.c. to maintain the desired radar i.f. will result in a deviation of the observed position of the local oscillator 'pip' from the centre of the pulse spectrum.

With the aid of such a device it is possible to check the locking accuracy of the a.f.c. under various conditions. Fig. 25 shows a typical picture obtained when an a.f.c. was locking correctly to a one-microsecond pulse, although some undesired ripple on the power supplies to the klystron was causing a small amount of frequency modulation of the local oscillator.

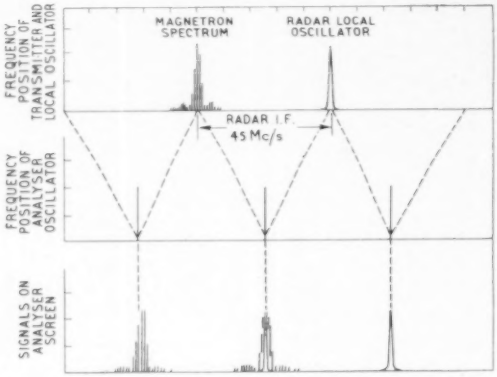
FREQUENCY POSITION OF TRANSMITTER AND LOCAL OSCILLATOR
MAGNETRON SPECTRUM
RADAR LOCAL OSCILLATOR
RADAR I.F. 45 Mc/s
FREQUENCY POSITION OF ANALYSER OSCILLATOR
SIGNALS ON ANALYSER SCREEN

Fig. 24. *Action of analyser in viewing signals.*

It will be appreciated that the i.f. of the spectrum analyser is being adopted as the standard of frequency, for the purpose of this test. It is, therefore, necessary to ensure high stability of this frequency. A double superheterodyne receiver was used in the spectrum analyser from which Fig. 25 was obtained. The first i.f. was 22·5 Mc/s and the final i.f. 2·5 Mc/s, a crystal-controlled oscillator being used to determine the remaining 20 Mc/s of the first i.f. so that the stability of the 22·5 Mc/s response could be relied on to 10 kc/s or so. Present experience suggests that such a device in skilled hands is a very valuable tool, but since it is as likely to go wrong as is the a.f.c. it is testing, it is emphatically not suitable for routine maintenance work.

The ideal and universally-applicable method of testing the transient response of an a.f.c. is to produce a step-function change in the transmitter frequency and to observe the rate at which the reflector voltage or tuning shaft responds. It is not easy to produce such a change in the frequency of a transmitter, however, and this test is only practicable if the transmitter sample to the a.f.c. mixer is replaced by the output of a suitable pulsed-signal generator. For example, Fig. 25 is

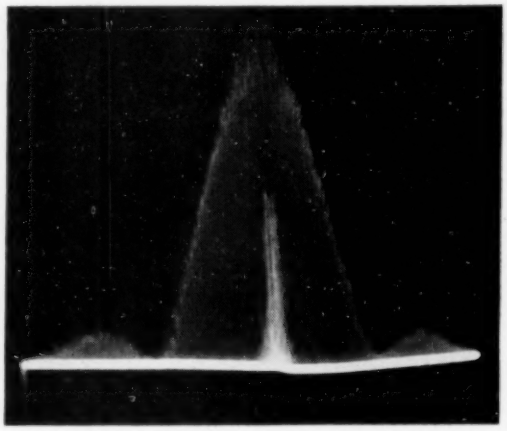

Fig. 25. *Typical oscillogram of the pulse spectrum with an a.f.c. system locking correctly on a 1-μsec pulse.*

a photograph of part of the spectrum of a one-microsecond X-band pulse produced by pulse-modulating the grid of a CV129 klystron. It is easy to apply a suitable square-wave to the reflector of this klystron and thereby frequency modulate the output pulse.

Despite the usefulness of such pulse signal generators as laboratory tools, this method of testing the a.f.c. performance is somewhat artificial, since it is difficult to ensure that the signal level at the a.f.c. mixer is that which obtains during normal operation. It is, therefore, desirable to have some way of testing the transient response of an a.f.c. system using as signal source the magnetron with which the a.f.c. is normally required to work.

One such method, which can often be applied, is to introduce a step-function disturbance in the reflector potential of the local oscillator. It is convenient to make this disturbance a periodic one, so that an oscilloscope can be used to observe the response. For example, if a resistance of a few thousand ohms is connected in series with the lead to the reflector, a high-speed relay may be arranged periodically to connect a $1\frac{1}{2}$-V battery, say, across this resistance. An oscilloscope can then be connected to display the reflector waveform, from which the transient response of the a.f.c. can readily be deduced. A useful refinement

is to brighten up the c.r.t. picture momentarily each time the transmitter fires, and to synchronize the switching frequency to some sub-multiple of the p.r.f.

Conclusion

A considerable amount of research has been devoted to improving the noise factor of centimetre-wave receivers. Much of the available improvement in performance can be lost by an unfortunate choice of receiver bandwidth and magnetron-pulling figure or by the use of an unsuitable a.f.c. system. This article has attempted to show the extent of the contribution which a good a.f.c. can make towards improving the overall performance of an equipment; to give enough circuit details to indicate the price that must be paid in increased numbers of valves and components; and to show that, despite the increased complexity, a good a.f.c. may simplify the maintenance problem by eliminating a number of preset adjustments, and thereby give both a better and more reliable performance.

It may reasonably be objected that it has been assumed that the equipment designer has complete control over the factors, such as the receiver bandwidth and magnetron-pulling figure, whereas in practice he is often compelled to use a particular i.f. amplifier or magnetron because of its availability. It is felt, however, that many systems are designed to use some particular components, not so much because better ones are not available, or cannot be developed in time, but because the designer is not aware of the extent of the loss in performance which he is tolerating for the sake of the convenience.

Experience has shown that by adopting the 'designable' a.f.c. circuit a better system can be obtained with less development work.

It is of some interest to compare the present-day problems of scanner pulling, frequency splitting, etc., with the problems of transmitter design in the early days of wireless telegraphy, when self-oscillatory output stages were in use. The latter problems included frequency shifts due to aerials swaying in the wind and varying in capacitance, and the 'Ziehen effect',[19] which showed itself as a spasmodic jump in transmitter frequency.

Automatic frequency control was invented to deal with these transmitter difficulties, but the solution finally adopted was to use a power amplifier at the output stage and to drive it from a separate master oscillator. It may be that eventually the same solution will be adopted in cm-wave radar systems, but much more progress in the technique of power amplification at these frequencies will be needed before this step becomes economical.

In the meantime, existing techniques are capable of dealing with the problem, at a price, and the main field in which improvement is to be expected is in the engineering details of the wide-range a.f.c. systems which are superseding the earlier systems.

Acknowledgments

The author is indebted to the Chief Scientist, Ministry of Supply, for permission to publish the results of this work, most of which was conducted under his auspices. The material has been drawn from the unpublished results of a large number of workers, but especial thanks are due to Mr. C. Baron of R.R.E., Malvern, for the circuit of Fig. 20, and for part of the discussion given in the Appendix.

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APPENDIX

Transient Response of an A.F.C. Loop

The behaviour of any servomechanism can be summed up in an 'Equation of Motion'. For a continuous feedback servo, this relates the position of the output shaft (or its equivalent) and, usually, derivatives of this quantity with respect to time, to the position of the input shaft (or equivalent), and possibly time derivatives of this quantity also.

The resulting differential equation can be solved, for a linear system, by straightforward, but possibly laborious, methods to give the response of the system to any applied disturbance, or to establish the stability conditions. For a linear sampling servo the same situation arises, except that the infinitesimal calculus must be replaced by the less familiar calculus of finite differences. Thus the equation of motion takes the form:

$$q_1 u_n + q_2 u_{n-1} + \ldots\ldots + q_n u_1 = f(n) \quad \ldots\ldots \quad (1)$$

where $u_1, u_2 \ldots\ldots u_n$ denote the values of the output quantity at the successive instants $t_1, t_2 \ldots\ldots t_n$ when samples are taken, and $f(n)$ is the applied disturbance.

It can be shown by substitution that equation (1) is satisfied by:

$$u_n = Aa^n + Bb^n + \ldots\ldots Nn^n + F(n) \quad \ldots\ldots \quad (2)$$

where $a, b, \ldots . n$, are the roots of the characteristic equation:

$$q_1 x^n + q_2 x^{n-1} + \ldots\ldots + q_n = 0 \quad \ldots\ldots \quad \ldots\ldots \quad (3)$$

and $A, B, \ldots . N$, are determined by the initial conditions and $F(n)$ gives the forced response to the input $f(n)$.

It follows from equation (2) that the stability condition for the sampling servo is that the modulus of each of the roots of equation (3), real or complex, should be less than unity. Just as the Laplace transform is commonly used by servo designers to simplify the solution of ordinary linear differential equations, integral transform techniques[21,22] are available for the solution of finite difference equations. For the reader who does not wish to master these techniques, we will present a simpler though more laborious treatment of the only two equations of serious interest to the a.f.c. designer, those relating to systems having respectively one and two significant time-lags.

Single Time-Lag

Consider the local oscillator and transmitter frequencies at the equally-spaced instants of time $t_0, t_1, \ldots . t_n$, etc.,

when the transmitter fires, and let the local-oscillator frequency at these instants be $f_0, f_1 \ldots f_n$, while the oscillator frequency required for correct tuning at these instants is $g_0, g_1 \ldots g_n$.

Then the tuning error e_n at time t_n is $(f_n - g_n)$, and the characteristic equation of the system is

$$f_n = f_{n-1} - Ae_{n-1} \quad .. \quad .. \quad .. \quad .. \quad (4)$$

where A is a constant determined by the circuit design.

Consider the response of the system to a step-function. Put $e_0 = E$ and $g_0 = g_1 = \ldots = g_n$. Then equation (4) becomes

$$e_n = E\,(1 - A)^n \quad .. \quad .. \quad .. \quad .. \quad .. \quad (5)$$

Thus the system is stable if $0 < A < 2$, and the response will be monotonic, 'dead-beat' or oscillatory, according as A is less than, equal to, or greater than unity.

Two Time-Lags

Using the above notation, suppose the output of the discriminator to be zero except for the time interval $t_0 < t < t_1$, when it then has a value which we may, without loss of generality, assume to be a constant e_0. Then during the interval $t_0 < t < t_1$:

$$\frac{df}{dt} = -Ke_0\left[1 - \exp -\left(\frac{t - t_0}{T}\right)\right]$$

where T is the speed build-up time-constant of the system.

$$f_0 - f_1 = Ke_0 \int_{t_0}^{t_1}\left[1 - \exp -\left(\frac{t - t_0}{T}\right)\right] dt$$

Putting $\dfrac{t_1 - t_0}{T} = r$,

$$f_0 - f_1 = e_0\,KT\,[r + \exp(\ -r) - 1] \quad .. \quad (6)$$

At any time $t_1 < t$

$$\frac{df}{dt} = -Ke_0\,[1 - \exp(\ -r)]\exp -\left(\frac{t - t_1}{T}\right)$$

whence $$f_1 - f_2 = Ke_0 \int_{t_1}^{t_2} [1 - \exp(\ -r)] \exp -\left(\frac{t - t_1}{T}\right) dt$$

Therefore $$f_1 - f_2 = e_0\,KT\,[1 - \exp(\ -r)]^2$$

and generally, for $n > 2$

$$f_{n-1} - f_n = e_0\,KT\,[1 - \exp(\ -r)]^2 \exp[-(n-2)\,r] \quad .. \quad .. \quad .. \quad (7)$$

Now if the output from the discriminator consists not of one signal e_0 but a sequence $\{e_n\}$, then the response of the oscillator frequency will be given by the sum of its responses to a series of inputs of the form just discussed. Thus equations (6) and (7), when applied successively to each term of $\{e_n\}$, give:

$$\begin{aligned} f_{n-1} - f_n = {} & e_{n-1}\,KT\,[r + \exp(\ -r) - 1] \\ & + e_{n-2}\,KT\,[1 - \exp(\ -r)]^2 \\ & + KT\,[1 - \exp(\ -r)]^2\,[\exp(\ -r)\,e_{n-3} \\ & + \exp(\ -2r)\,e_{n-4} + \text{etc.}] \end{aligned}$$

With the aid of a similar expression for $(f_{n-2} - f_{n-1})$, the infinite series can be eliminated, giving:

$$\begin{aligned} f_n - f_{n-1}\,[1 + \exp(\ -r)] + f_{n-2}\exp(\ -r) \\ = -e_{n-1}\,KT\,[r + \exp(\ -r) - 1] \\ - e_{n-2}\,KT\,[1 - \exp(\ -r)]^2 \\ + e_{n-2}\,KT\,[r + \exp(\ -r) - 1]\exp(\ -r) \end{aligned} \quad .. \quad (8)$$

To find the response of the system to a step function, as before, put $g_0 = g_1 = \ldots = g_n = G$, say. Then equation (8) leads to:

$$\begin{aligned} e_n - \{1 + \exp(\ -r) - KT\,[r - 1 + \exp(\ -r)]\}\,e_{n-1} \\ + \{\exp(\ -r) + KT\,[1 - (1 + r)\exp(\ -r)]\}\,e_{n-2} \\ = 0 \end{aligned} \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (9)$$

It is then possible to compute the response to a step E by using:

$$e_1 = E\,\{1 - KT\,[r + \exp(\ -r) - 1]\}$$
$$e_2 = e_1 - KT\,[1 - \exp(\ -r)]^2 E$$

and equation (9) for the later terms.

Alternatively, writing equation (9) as:

$$e_n + Ae_{n-1} + Be_{n-2} = 0 \quad .. \quad .. \quad .. \quad (10)$$

it is possible to use equation (2) for terms later than the second. Since, in deriving equation (9), it was assumed that the transmitter frequency had remained constant for at least two pulses, this equation is not valid for e_1 and e_2. The stability conditions for the system are that the roots of the characteristic equation should have a modulus less than unity. These are satisfied if:

$$f(1) = 1 + A + B > 0$$
$$f(-1) = 1 - A + B > 0$$
$$f(0) = B < 0$$

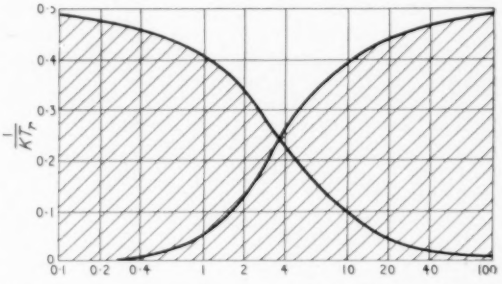

Fig. 26. *Stability conditions for system with two time-lags.*

These conditions have been used to plot the curves of Fig. 26. The shaded area shows the unstable region.

The interested reader should compare this analysis with a different approach due to Hurewicz.[22]

Effect of Component Tolerances on the Performance of an Extrapolating A.F.C. Loop

Consider the single time-lag system discussed above. Equation (5) gives the response to a step-function change in transmitter frequency as

$$e_n = E\,(1 - A)^n$$

where e_n denotes the tuning error n pulses after the application of the step of height E, and A denotes the ratio of the amount the oscillator frequency moves between one pulse and the next, to the tuning error during the corresponding period. The best performance will be attained when $A = 1$, but some deviation from this value must obviously occur in practice, and it was shown previously that A may well vary over a range of 10 : 1. It is now proposed to derive an expression showing the effect on performance of given tolerances on A.

As a measure of the efficiency of the system, consider the number of pulses necessary to reduce e_n/E to some amount $1/K$.

From equation (5)

$$1/K = (1 - A)^n$$

$$\text{or} \quad n = \frac{\log 1/K}{\log (1-A)} = \frac{-\log K}{\log (1-A)} \qquad (11)$$

It is necessary to choose a, the geometric-mean value of A. This will be chosen to keep the maximum value of n as low as possible, by making the performance fall off equally at both limits of A.

Let these limits be ar, a/r where $r < 1$.

Since we are only concerned with the modulus of the error, the desired condition is given by putting

$$|1 - ar| = \left|1 - \frac{a}{r}\right|$$

$$\text{or} \quad a = \frac{2r}{r^2 + 1} \qquad (12)$$

Substituting ar for A in equation (11):

$$n = \frac{-\log K}{\log \dfrac{1 - r^2}{1 + r^2}}$$

Since $r^2 < 1$

$$= \frac{\log_e K}{-2\left(r^2 - \dfrac{r^6}{3} - \dfrac{r^{10}}{5}\right)}$$

If $r^2 < 0{\cdot}5$, we may neglect the terms in r^6, r^{10}, etc., without introducing a serious error. Thus we may write

$$n = \frac{\log_e S}{2r^2}$$

$$\text{or, writing} \quad R = \frac{1}{r^2}, \; n = \frac{R \log_e K}{2} \qquad (13)$$

where R is the ratio of the greatest possible value of A to the smallest. In other words, the system has a time-constant equal to, or less than, $\frac{1}{2}RT$, where:

$$T = (t_n - t_{n-1}) = \frac{1}{\text{p.r.f.}}$$

It is clear from equation (13) that a good transient response can only be guaranteed if R can be kept low.

Similar results hold for systems having more than one time-lag, since, although it is possible to devise circuits which to some extent compensate for variations in A when following a smoothly-varying transmitter frequency, it has already been shown that this problem is not that with which we are concerned.

'Instantaneous' A.F.C. System

This system, discussed previously, attempts to avoid the fundamental limitation discussed above by arranging that the a.f.c. behaves, during the pulse, as a normal negative-feedback system.

Baron's circuit, shown in Fig. 20, has a peak separation sufficiently wide to make it possible to neglect the phase shift in the i.f. amplifier. It may be assumed that there are, in all, three significant time-lags: in the diode loads, in the 'clamp' circuit, and in the anode load of the subsequent d.c. amplifier. It may be shown that the most economical arrangement makes the time-constant of two of these circuits equal to each other, while the third must be much larger in order to ensure stability. Let the long and short time-constants be t_1 and t_2 respectively, and let the gain round the feedback loop be μ. Then if F_t is the frequency error at time t, and if F' is the frequency change produced at the local oscillator, the 'equation of motion' is given by $(1 + t_1 D)(1 + t_2 D)^2 F' = \mu F_t$.

Since $t_1 \gg t_2$, the auxiliary equation may be simplified approximately to:

$$p^3 + \frac{2}{t_2} p^2 + \frac{1}{t_2^2} p + \frac{\mu}{t_1 t_2^2} = 0$$

Putting $p = 1/t_2 \, P$ (i.e., changing the time scale so that $t = t_2 T$) we obtain:

$$P^3 + 2P^2 + P + \mu t_2/t_1 = 0 \qquad (14)$$

This equation may be solved graphically by plotting $f(P) = P^3 + 2P^2 + P$ in the range $0 > P > -2$ (this being the range within which the system is stable). The solution for any value of $\mu t_2/t_1$ is then obtained direct from the curve, since $\mu t_2/t_1 = -f(P)$. The system is critically damped for $\mu t_2/t_1 = 0{\cdot}14$, and the response to a unit step function for this value and other values of $\mu t_2/t_1$ are shown in Fig. 27, which is due to C. Baron, as is an analysis from which the above treatment is derived.

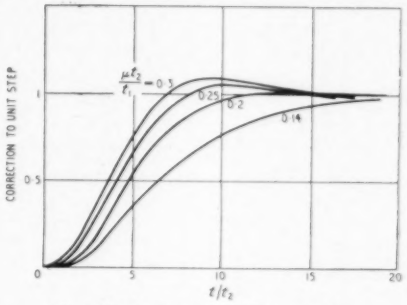

Fig. 27. *Response of 'Baron' a.f.c.*

Effect of Component Tolerances on the Performance of an 'Instantaneous' A.F.C. Loop

Since the main objection to 'extrapolating' a.f.c. systems is the considerable departures from the ideal performance which are occasioned by the tolerances on the gain round the feedback loop, it is of some interest to consider the effect of these tolerances on the system discussed above.

We first require an approximation to the response of the system for values of $\mu t_2/t_1$ appreciably higher or lower than the optimum value. For low values the response consists of three exponential terms, and it will be assumed here that only the slowest of these is important. Thus it will be assumed that the response is determined by the smallest root of equation (14). For values of $\mu t_2/t_1$ greater than 0·14, the equation (14) has only one real root and the response is oscillatory. It is proposed to consider the response as determined by the envelope of this oscillation; i.e., by the real part of the complex roots.

Equation (14) can readily be solved graphically, as indicated above. The value of the real part of the complex roots, when these occur, can be found when the real root, $-r$ say, is known, by using the relation that the sum of the three roots is minus one times the coefficient of x^2 in equation (14). It follows that the real part of the complex roots is given by:

$$a = \frac{r - 2}{2} \qquad (15)$$

Fig. 28 then gives a value of p giving a suitable approximation to the response of the system to a step function of unit height in the form

$$C_t = 1 - \exp\left(-p \frac{t}{t_2}\right) \qquad (16)$$

where C_t is the error at time t after the application of the step. It will be remembered that, for values of $\mu t_2/t_1$ corresponding to the upper part of the curve, equation (16) gives the envelope only of the oscillation.

Suppose now that the tolerances on the gain round the

loop can be estimated, and that the ratio of the maximum possible gain round the loop to the minimum is R say, where a probable value for R is about 10. It is now necessary to choose the mean value for $\mu t_2/t_1$ which gives the best performance over the range R. Since we are only concerned with the modulus of the error, it will be

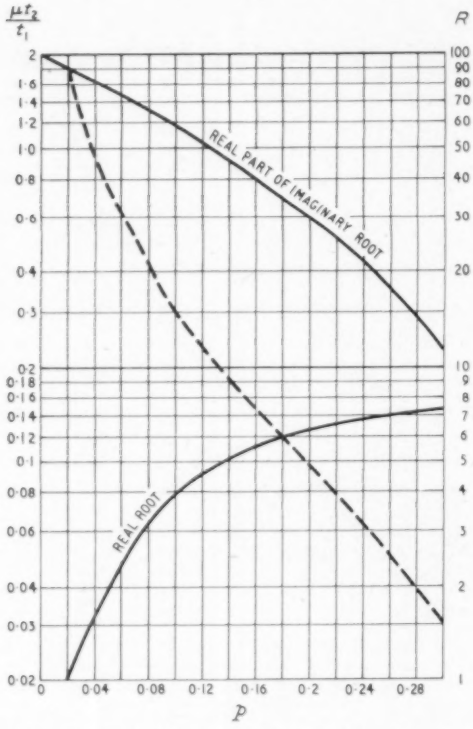

Fig. 28. *Smallest real root or real part of complex root of* $p^3 + 2p^2 + p + \mu t_2/t_1 = 0$.

sufficient to choose the mean value such that the two extreme tolerances give the same value of p in Fig. 28, for clearly if any other mean is chosen, the 'guaranteeable' performance will deteriorate. The dotted curve gives the ratio R, of the two values of $\mu t_2/t_1$ for each value of p.

The response time-constant p must be chosen to suit the pulse width and the extent to which it is proposed to attempt to correct the tuning error during a given pulse.

It will be seen from Fig. 27 that, although the Baron a.f.c. makes possible a more rapid response for given tolerances, these tolerances have a similar effect on the response speed to that which they have on the response of an 'extrapolating' servo.

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STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for May 1954

Date 1954 May	Frequency deviation from nominal: parts in 10^9		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	— 1·3	+ 4	NM
2	— 1·3	+ 4	NM
3	— 1·2	+ 5	NM
4	— 1·2	+ 4	+ 6·5
5	— 1·2	+ 4	+ 5·0
6	— 1·3	+ 4	+ 4·2
7	— 1·2	+ 4	+ 2·5
8	— 1·2	+ 3	NM
9	— 1·1	+ 4	NM
10	— 1·2	+ 4	— 1·6
11	— 1·1	+ 3	— 2·6
12	— 1·0	+ 4	— 3·6
13	— 1·1	+ 3	— 4·8
14	— 1·3	+ 2	— 5·9
15	— 1·3	+ 4	NM
16	— 1·3	+ 5	NM
17	— 1·3	+ 4	— 8·5
18	— 1·3	+ 4	— 9·2
19	— 1·2	+ 4	— 10·6
20	— 1·2	+ 4	— 11·0
21	NM	+ 4	— 11·4
22	— 1·2	+ 4	NM
23	— 1·3	+ 4	NM
24	— 1·2	+ 5	— 12·8
25	— 1·2	+ 5	— 12·6
26	— 1·2	+ 4	— 13·8
27	— 1·1	+ 6	— 14·4
28	— 1·1	— 1	— 14·2
29	— 1·1	— 2	NM
30	NM	— 2	NM
31	— 1·1	— 2	— 13·4

The values are based on astronomical data available on 1st June 1954.
NM = Not Measured.

NEW BOOKS

Statistical Methods in Electrical Engineering

By D. A. Bell. Pp. 175. Chapman & Hall, 37 Essex Street, London, W.C.2. Price 25s.

Some knowledge of statistics is almost indispensable for scientists, research workers and 'technologists'. Unfortunately, from the point of view of the majority of these workers, statistics is an essentially mathematical subject. From the point of view of a mathematician, like the reviewer, statistics is best treated as a branch of mathematics. There is a certain amount of fundamental technique and theory which must be mastered whatever the application; this includes the practical estimation of the mean, variance, and other moments, of correlation coefficients, and the theory of various standard statistical distributions and the relations between them. But there are some who, perhaps rightly from their own point of view, regard mathematics as a necessary evil which they wish to avoid as far as possible. If they are experienced engineers with little previous knowledge of statistics, they will find Dr. Bell's book helpful. For though Dr. Bell's style is somewhat involved, and there are too many confusing parenthetic insertions and asterisked explanatory footnotes, he has written in a language readily understood by experienced engineers. The fundamental technique and theory mentioned above is covered, but the distinctive part of Dr. Bell's book is the relation of statistics to such subjects as waveform analysis, noise and information theory.

The book arose from a series of lectures to post-graduate students of Birmingham University: it is not intended for the undergraduate student, for whom the language familiar to electrical engineers may require as much learning as the language of mathematics. There should not be the present unnatural separation between the subjects of radio-electricity and statistics, which are in fact closely related. Dr. Bell's book is welcome because it helps to reduce this separation. J. W. H.

Prediction of Optimum Traffic Frequencies

By Ing. Prof. Rufino Gea Sacasa. Pp. 20 with 2 diagrams and 5 nomograms. Obtainable from the author, El Encimar 10, Madrid, Spain. Price 38s.

For some years the production of ionospheric forecasts has been an important ancillary service to organizations engaged in radio communication. The laboratories responsible for preparing the forecasts usually issue six charts to represent conditions in the F_2 layer during a given month at a particular epoch in the solar cycle. These charts already contain approximations and it would be difficult to reduce their number without a serious loss of accuracy. In spite of the fact that the use of such charts is complicated, the tendency on the part of the operational users is to demand higher accuracy rather than a simpler means of presenting the data.

Prof. Gea Sacasa's method represents an attempt to compress into five nomograms all the information required to make forecasts of the optimum traffic frequency (f.o.t.) in all parts of the world at any part of the solar cycle and for any desired month. To achieve this, a very drastic simplification of what does in fact occur is inevitable and it is worth pointing out the implications which are inherent in the use of the nomograms.

The basic assumption which is made, referred to as the 'probable law', is that the f.o.t. for a 1300-km NS path is a linear function of time of day and that it increases until 1400 L.M.T. and falls thereafter. The functions used are: Morning: f.o.t. $= 10 + 2n$ Mc/s; Evening: f.o.t. $= 18 - 2n$ Mc/s where n = number of hours after sunrise or sunset as appropriate. From the way in which these relations are used, it is possible to derive from them the implied relations for the critical frequencies (f_0E, f_0F_2) of E and F_2 layers:

Morning: $f_0E = 2{\cdot}2 + 0{\cdot}5n$ Mc/s; $f_0F_2 = 3{\cdot}9 + 0{\cdot}8n$ Mc/s

Evening: $f_0E = 4{\cdot}0 - 0{\cdot}5n$ Mc/s; $f_0F_2 = 7{\cdot}0 - 0{\cdot}8n$ Mc/s

A representative sample of actual values of f_0F_2 and its slope at sunrise and sunset has been taken for 1952; the mean values of these parameters and their standard deviations lead to the relations:

Morning: $f_0F_2 = 3{\cdot}5 \pm 0{\cdot}7 + n(1{\cdot}2 \pm 0{\cdot}7)$ Mc/s

Evening: $f_0F_2 = 5{\cdot}9 \pm 1{\cdot}8 - n(1{\cdot}0 \pm 0{\cdot}9)$ Mc/s

It seems likely that the differences between the observed relations and those implied by the 'probable law' could be explained in terms of different levels of solar activity. For this reason it is difficult to see how any single 'probable law' can be satisfactory throughout the solar cycle. Quite apart from this criticism of the method, potential users of the nomograms must satisfy themselves that their needs can be met by a triangular wave approximation for the diurnal changes in f_0E and f_0F_2; and by the substitution of mean values for parameters whose standard deviations vary from 20% to 90% of the mean value.

Although a compact nomographic technique for calculating approximate values of f.o.t. would be useful, it seems fair to suggest that Prof. Gea Sacasa may have simplified his approximations to such an extent as to limit rather severely possible applications of his method.

C. M. M.

Thermionic Valves

By A. H. W. Beck, B.Sc.(Eng.), A.M.I.E.E. Pp. 570 + xvi. Cambridge University Press, 200 Euston Road, London, N.W.1. Price 60s.

This book is in three parts. The first covers the physical theory of electronics in three chapters on thermionic emission, secondary, field and photoelectric emission, and fluorescence and phosphorescence. The second provides the mathematical theory of electronics and covers electrostatic fields, electrostatic and magnetic electron optics of fields without space charge, space-charge flow and the diode, transit-time effects and fluctuation noise in valves. In the third part, types of valve are discussed and there is one chapter on triodes for low and medium frequencies, another on multigrid receiving valves and a third on transmitter valves. The interest then shifts to the higher frequencies in chapters covering velocity-modulated valves, triodes at ultra-high frequencies, travelling-wave tubes and beam interaction tubes, magnetrons and picture convertors and storage tubes.

The book is intended mainly for "graduates with a first degree in physics or electrical engineering and those who are starting independent work in this field, either in industry or in the various post-graduate courses on electronics".

Mathematics are freely used throughout, but not for their own sake. There are many consecutive pages of reading matter with no sign of an equation. It is, however, true to say that a good mathematical knowledge is necessary to follow a great deal of the discussion.

Ordinary types of valve are treated somewhat briefly, but not cathode-ray tubes nor gas-filled types. The whole emphasis is on valves for the higher frequencies.

Radio Research 1953

Report of the Radio Research Board for 1953. Pp. 40 + iv. Published for D.S.I.R. by H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 1s. 9d.

Laplace Transforms for Electrical Engineers

By B. J. STARKEY, Dipl.Ing., A.M.I.E.E. Pp. 279. Published for *Wireless Engineer* by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 30s.

In the first two chapters, this book deals with the symbolic method, Fourier transformations, generalized impedances and cisoidal oscillations and, in Chapter 3, introduces the Laplace transform. Its application to electrical circuits comes in Chapter 4 and the next four chapters cover functions of a complex variable, integration in the complex plane, classification of functions of a complex variable and further analysis of contour integration. Chapters on Laplace transformation, its applications and theorems, the use of the inverse Laplace transformation and the generalization of the theory follow.

The book is intended for engineers rather than for the mathematician and the author claims to have used a physical rather than a mathematical vocabulary.

Applied Electronics Annual 1953-1954

Edited by R. E. BLAISE, A.M.Brit.I.R.E. Pp. 257. British-Continental Trade Press, Ltd., 222 Strand, London, W.C.2. Price 20s.

Discrete Sources of Extra-Terrestrial Radio Noise

Special Report No. 3 of the International Scientific Radio Union. Pp. 56. Available from the General Secretariat of the U.R.S.I., 42 Rue des Minimes, Brussels, Belgium. Price 75 fr.

Index to "Proc.I.E.E.", 1942-51

A 10-year index to the *Proceedings of the Institution of Electrical Engineers*, covering Parts I, IA, II, IIA, III, IIIA and Monographs (i.e., Part IV). It includes author and subject indexes. Pp. 500. The Secretary, The Institution of Electrical Engineers, Savoy Place, London, W.C.2. Price 25s.

Handbook of Industrial Electroplating (2nd Edition)

By E. A. OLLARD, A.R.C.S., F.R.I.C., F.I.M., and E. B. SMITH. Pp. 364. Published for *Metal Industry*, by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 30s.

The Oscilloscope at Work

By A. HAAS and R. W. HALLOWS, M.A.(Cantab.), M.I.E.E. Pp. 171 with 102 diagrams and 217 oscillograms. Published for *Wireless World* by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 15s.

This book was originally published in France. It has been translated and enlarged, and covers the form and applications of the general-purpose oscilloscope.

OBITUARY

It is with deep regret that we have to announce the death on 21st June of Francis Morley Colebrook after a short illness. Born in 1893, Mr. Colebrook was educated at Birkbeck College, London, and the City and Guilds College. He was with the Radio Division of the National Physical Laboratory from 1926-1949 when he became Officer-in-Charge of the Electronics Section.

He has contributed to *Wireless Engineer* and *Wireless World* and was the author of "Basic Mathematics for Radio Students". Since 1946 he has been a member of the *Wireless Engineer* Editorial Advisory Board.

As announced elsewhere, he was appointed an Officer of the Order of the British Empire in this year's Birthday Honours.

NATIONAL BUREAU OF STANDARDS:

Statistical Theory of Extreme Values and Some Practical Applications

By E. J. GUMBEL. National Bureau of Standards Applied Mathematics Series 33. Pp. 51. Price 40 cents.

Optical Image Evaluation

Proceedings of Symposium held October 1951. National Bureau of Standards Circular 526. Pp. 289. Price $2.25.

Tables of Lagrangian Coefficients for Sexagesimal Interpolation

National Bureau of Standards Applied Mathematics Series 35. Pp. 157. Price $2.

Table of Secants and Cosecants to Nine Significant Figures at Hundredths of a Degree

National Bureau of Standards Applied Mathematics Series 40. Pp. 46. Price 35 cents.

Effective Radio Ground-Conductivity Measurements in the United States

By R. S. KIRBY, J. C. HARMAN, F. M. CAPPS and R. N. JONES. National Bureau of Standards Circular 546. Pp. 87. Price 65 cents.

The National Bureau of Standards' publications can be obtained from the Government Printing Office, Washington 25, D.C., U.S.A. One-third of the publication price should be added to cover postage.

BIRTHDAY HONOURS

In the Birthday Honours List, Major-General L. B. Nicholls, C.B., C.B.E., M.I.E.E. (Chairman of Cable & Wireless) is appointed K.C.M.G.

T. E. Goldup, M.I.E.E. (Director, Mullard, Ltd., and Chairman of the Board of Governors of the Ministry of Supply School of Electronics at Malvern), has been appointed a Commander of the Order of the British Empire.

F. M. Colebrook, B.Sc., A.C.G.I., D.I.C. (Officer in charge of the Electronics Division, National Physical Laboratory), becomes an Officer of the Order of the British Empire.

A. I. Forbes Simpson of the radio equipment design department in the Coventry works of the General Electric Co., Ltd. becomes a Member of the Order of the British Empire.

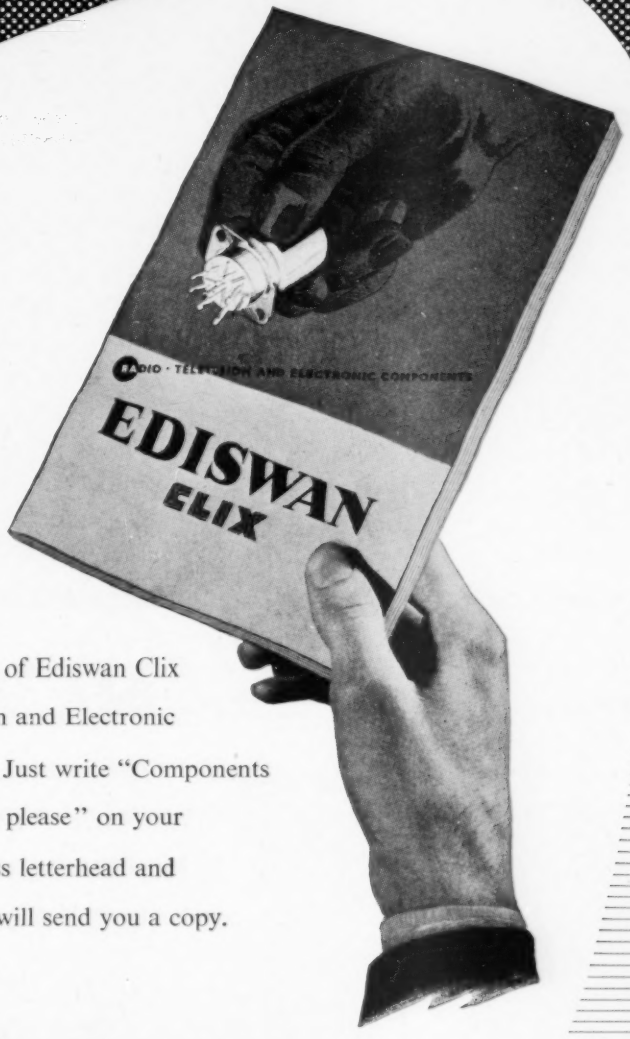
INSTITUTE OF PHYSICS

Sir John Cockcroft (Director of the Atomic Energy Research Establishment, Harwell) was recently elected President of the Institute of Physics. G. R. Noakes (Assistant Master, Uppingham School) was elected Vice-President. Dr. S. Whitehead and Dr. B. P. Dudding were re-elected Honorary Treasurer and Honorary Secretary respectively.

INSTITUTION OF ELECTRONICS EXHIBITION

The ninth annual Electronics Exhibition is being held at the Manchester College of Technology, Sackville Street, Manchester, 1, from 14th-20th July. It will be open on Wednesday, 14th July, from noon to 10 p.m. and on other days (except Sunday) from 10 a.m. to 10 p.m. except Saturday, 17th July, when it will close at 7 p.m. The exhibition will be in two sections, one covering scientific and industrial research items and the other commercial apparatus.

Tickets are obtainable from W. Birtwistle, 78 Shaw Road, Thornham, Rochdale, Lancs.



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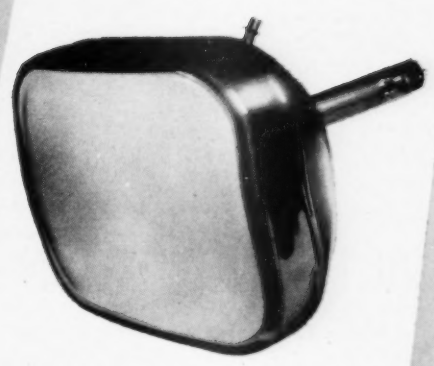

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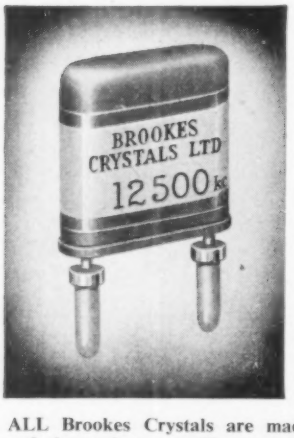

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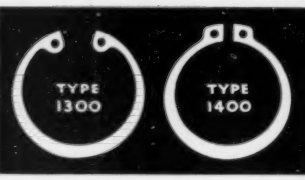

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Ministry of Transport and Civil Aviation: Telecommunications Technical Officers, Grade III. By arrangement with the Civil Service Commissioners applications are invited from men for at least 12 posts. Age at least 23 on 1st June, 1954. Candidates must possess (*a*) the Ordinary National Certificate in Electrical Engineering or (*b*) City and Guilds Intermediate Certificate in Telecommunications Engineering, plus City and Guilds Certificate Radio III or (*c*) City and Guilds Certificates Telecommunications Principles III (or higher) and Radio III (or higher) or (*d*) produce evidence of an equivalent standard of technical education. They must also have had not less than 8 years' experience in an appropriate telecommunications field, including manufacturing or maintenance or operation.

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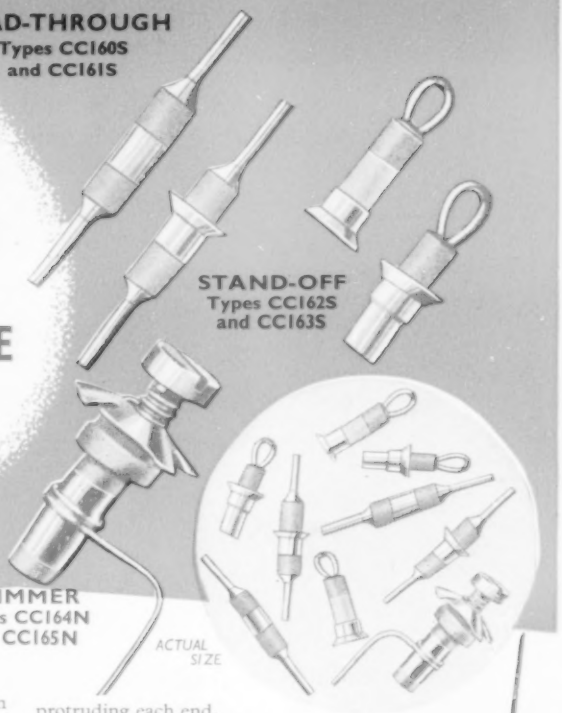

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Capacity sweeps supplied: 0.5 pF. to 3 pF., and 1 pF. to 5 pF. Temperature Coefficient: −40 10^6/°C. Working Voltage: 350v D.C. Insulation Resistance: greater than 10,000 megohms at 350v. D.C.

CONSTRUCTION

Lead-Through and Stand-Off: the tubular bodies are a specially selected material of the barium-titanite group of ceramics with a dielectric constant of 3,000. The four constructions provide two for lead-through, and two for stand-off purposes. The former have an 18 gauge tinned copper wire running through the body and protruding each end, whilst the latter have a looped termination at one end.

Trimmers: the 'live' electrode is a silvered band on the outside of the tube, contact being provided by a tinned copper wire termination. Variation in capacity is controlled by a screw in an internally threaded ceramic tube.

INSTALLATION

Lead-Through Type CC160S is for soldering direct into a tuner plate, through a 'burst' hole.

Lead-Through Type CC161S has a central flange so that it may be mounted on a tuner plate having a punched hole.

Stand-off Type CC162S has a mounting flange at the base enabling the condenser to be soldered to any part of a flat chassis.

Stand-off Type CC163S has a central mounting flange, and is mounted through a punched hole in the chassis.

Trimmer Types CC164N & CC165N. The square-section end is inserted in a suitably punched hole, thus preventing movement of the body during adjustment. The condenser is held by a spring fixing nut which locks the adjusting screw, and gives a low inductance connection to chassis.

TCC

THE TELEGRAPH CONDENSER CO. LTD

RADIO DIVISION

NORTH ACTON · LONDON · W3 · Tel: ACOrn 0061

SPECIALISTS IN CONDENSERS SINCE 1906

Approved to ENTIRELY NEW Inter-Service Standards

A miniature range of Hunts Electrolytics approved to Inter-Service standards RCS 134 Issue 3 and RCL 134 Addendum Issue 2.

Temperature range: −30° C to +70° C max.

Please write for leaflet giving details of complete range.

Cap µF	Peak Working Volts at 70°C.	Max. R.M.S. Current at 50/c/s (mA)	Dimensions (inches) L.	D.	List Number	Inter-Service Cat. Number
TYPE L31/I. PATTERN CE4. CLASS HI						
50	25	70	1¼	⅜	JB 53AKZ	Z145512
100	25	100	1¾	⅜	JB 54KZ	Z145514
1000	25	600	3	1	JB 57KZ	Z145520
25	50	60	1¼	⅜	JB102BKZ	Z145508
50	50	100	1¾	⅜	JB103KZ	Z145513
500	50	450	3	1	JB106AKZ	Z145519
8	150	60	1¼	⅜	JB153BKZ	Z145502
16	150	90	1¾	⅜	JB154KZ	Z145505
32	150	160	1¾	½	JB181KZ	Z145509
8	350	75	1¾	⅜	JB403KZ	Z145503
16	350	120	1¾	½	JB405KZ	Z145506
32	350	225	2	1	JB407AKZ	Z145510
4	450	50	1¾	⅜	JB552KZ	Z145501
8	450	100	1¾	½	JB553BKZ	Z145504
16	450	175	2	1	JB554AKZ	Z145507
32	450	275	3	1	JB555AKZ	Z145511
TYPE L32/I. PATTERN CE5 CLASS HI						
3000	25	1100	4½	1⅜	KB 62KZ	Z145557
1500	50	1000	4½	1⅜	KB111KZ	Z145555
60	350	350	2	1⅜	KB430KZ	Z145552
100	350	450	3	1⅜	KB411KZ	Z145554
32	450	275	3	1	KB555BKZ	Z145551
60	450	450	3	1⅜	KB581KZ	Z145553
TYPE L32/3. PATTERN CE6. CLASS HI						
32+32	350	200	2	1⅜	KB417KZ	Z145601
60+100	350	400	4½	1⅜	KB420KZ	Z145603
60+250	350	400	4½	1¾	KB422KZ	Z145605
100+200	350	550	4½	1¾	KB423KZ	Z145606
32+32	450	300	3	1⅜	KB564AKZ	Z145602
60+100	450	550	4½	1¾	KB565KZ	Z145604

REGISTERED TRADE MARK

HUNTS CAPACITORS

THE TRADE MARK OF RELIABILITY

In accordance with Government specification all units are supplied insulated by an approved sleeve, the dimensions being increased accordingly by ⅛" on the length and 1/16" on the diameter.

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